WIRELESS ENGINEER

THE JOURNAL OF RADIO RESEARCH & PROGRESS

JANUARY 1954

VOL. 31 No. 1 THREE SHILLINGS AND SIXPENCE

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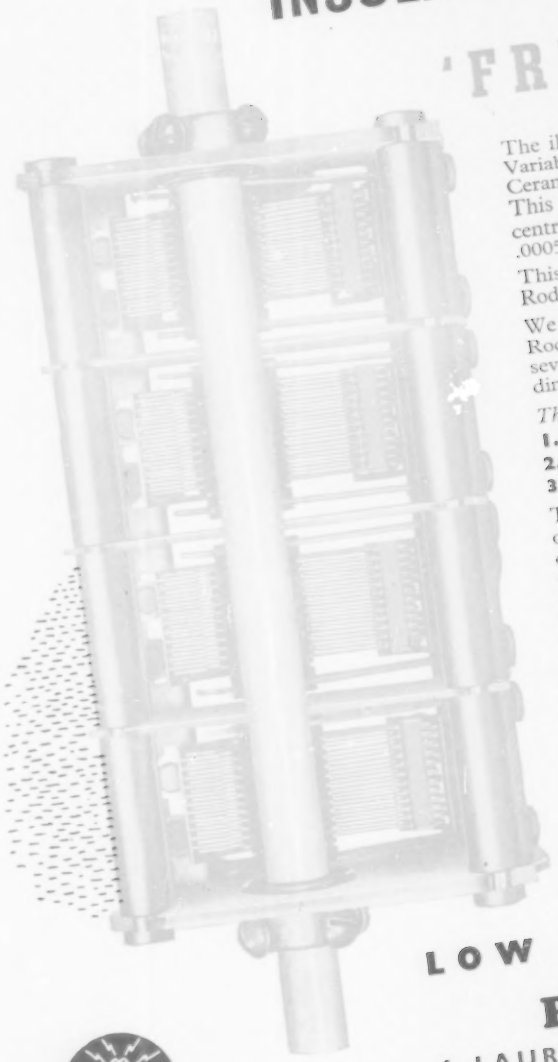

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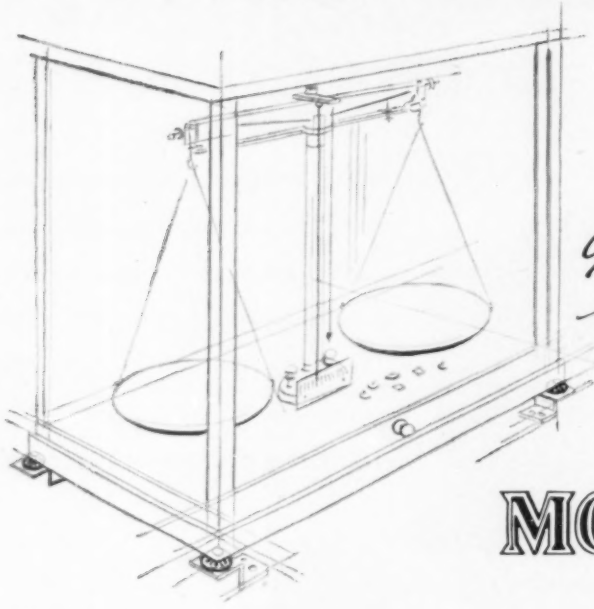

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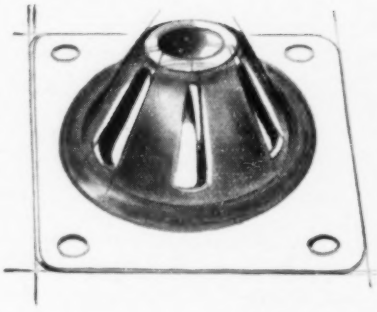

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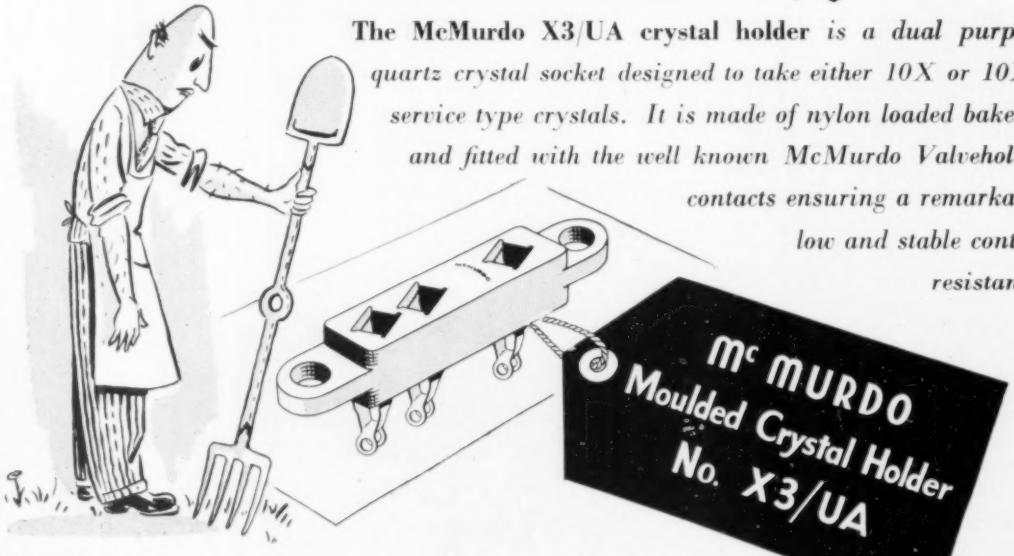

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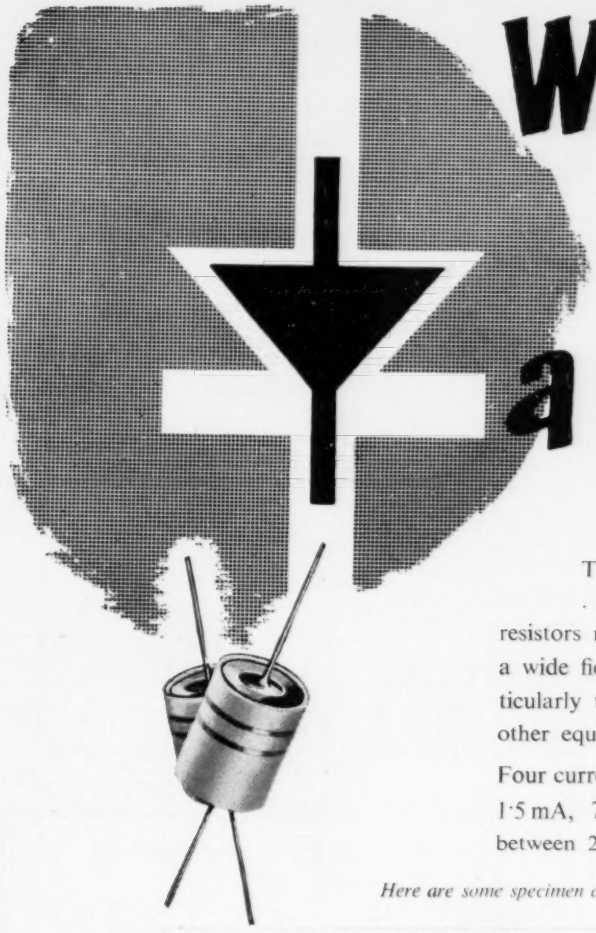

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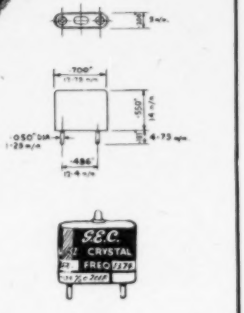

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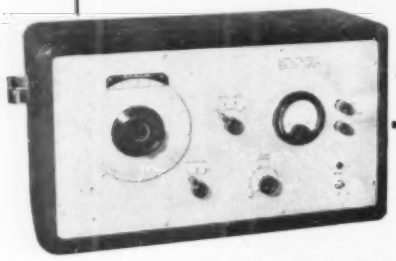

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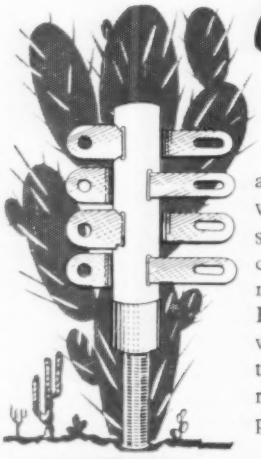

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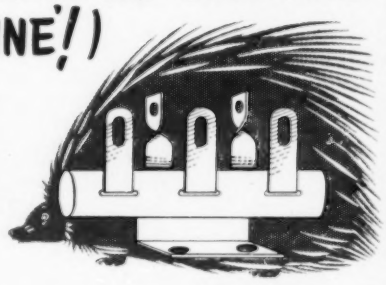

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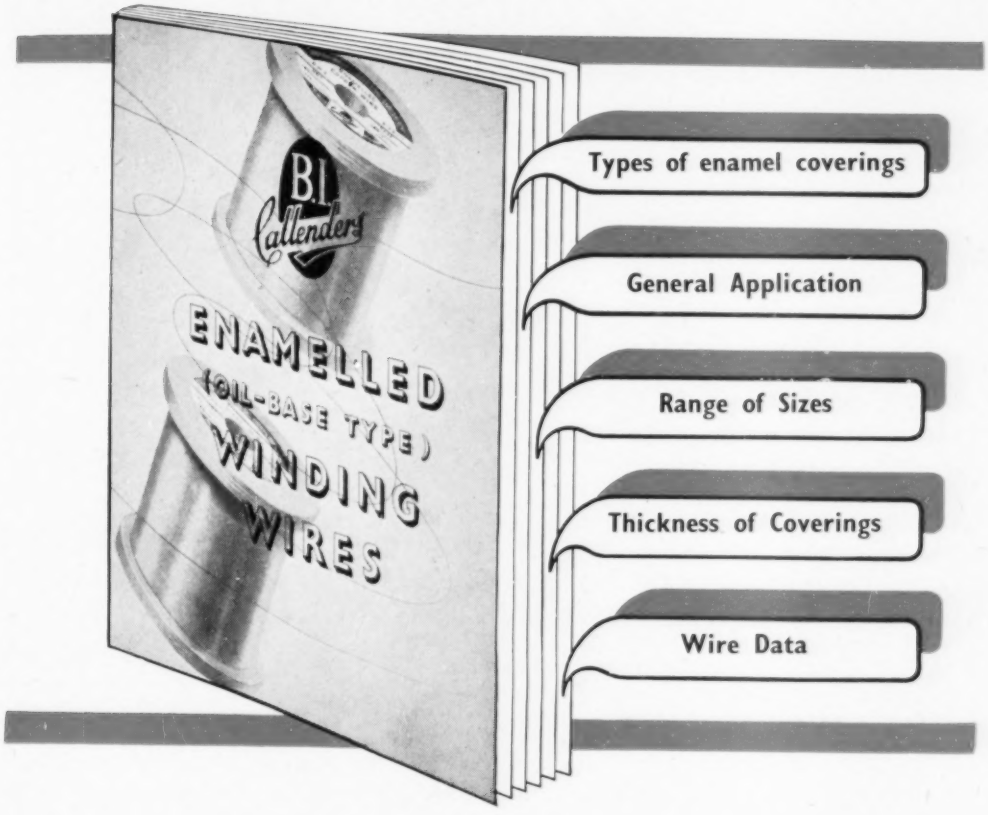

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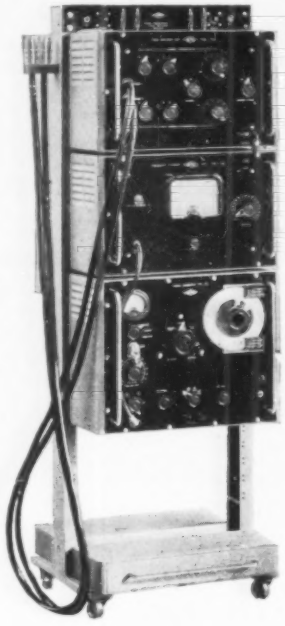

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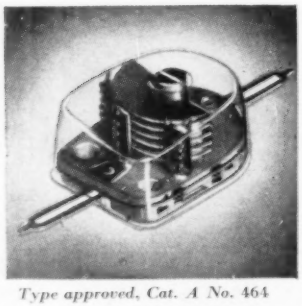

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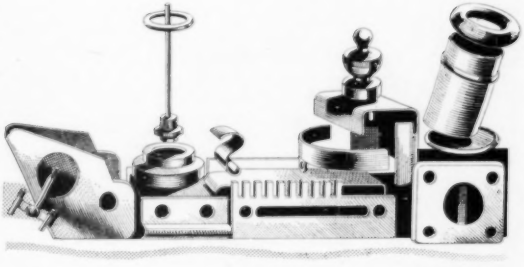

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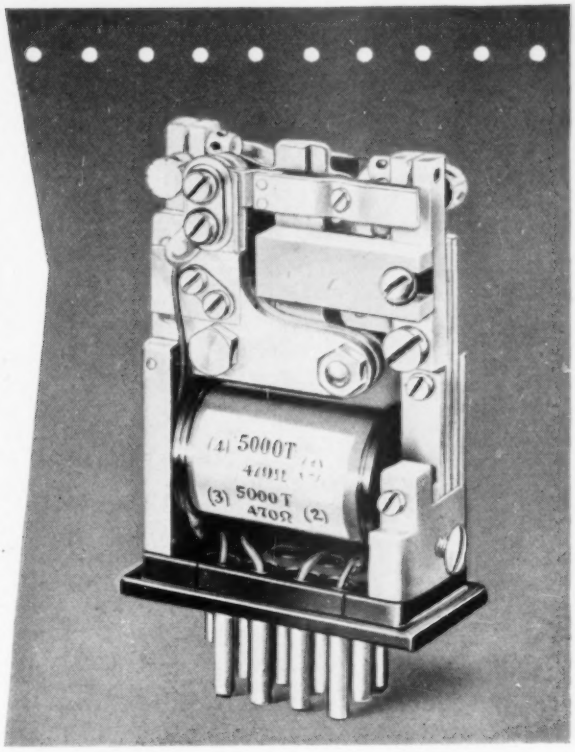

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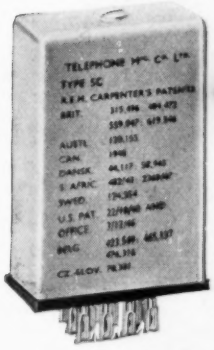

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Volume 31 · Number 1

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MVT.146

WIRELESS ENGINEER

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Ferrite-Core Inductors

IN the October Editorial we discussed the developments that had taken place in sintered magnetic materials suitable for use as cores at high frequencies. Although preliminary experiments were made over 40 years ago, it was the work of Snoek, in the Philips Laboratories in Holland, about 20 years ago that led to the great developments that have since taken place. A recent article by H. A. Stone in the *Bell System Technical Journal* (March 1953) with the above title deals in a very interesting manner with the design problems that arise if one wishes to take full advantage of the properties of the ferrite material.

The alternative to ferrite is powdered molybdenum permalloy, which has its limitations and disadvantages. The component has to be pressed into the required shape and is fragile; the higher the frequency the smaller must be the size of the particles, and this not only introduces mechanical difficulties but reduces the effective permeability since more of the space is occupied by the binding insulating material. The replacement of the powdered cores by ferrites will necessitate a reconsideration of the design.

If, instead of Q, one uses its reciprocal, one can write

$$D = \frac{1}{Q} = \frac{R_{dc} + R_e + R_h + R_r + R_c + R_s}{\omega L}$$

where D is the total dissipation factor,

R_{dc}, the d.c. resistance of the winding,

R_e, R_h and R_r, the effective resistances due to eddy currents, hysteresis and residual losses respectively,

R_c, the increase of resistance due to distributed capacitance and,

R_s, the increase in the resistance of the wire due to the alternating current.

For specified values of L, I and f and for a given core material each of the above will depend on the effective permeability, volume, and shape of the core, and, in the article referred to, these three factors are explored with the object of getting the lowest possible value of D. With ferrites the effective permeability μ is easily adjusted by varying the air-gap. The first four of the above six items of loss are direct functions of the permeability. If A is the cross-sectional area and l the length of path, then $L = kN^2A\mu/l$ where k is a constant and N the number of turns. If the shape is fixed and V is the volume of the core then $A/l = V^{\frac{2}{3}}/V^{\frac{1}{3}} = V^{\frac{1}{3}}$ and $L = k_1N^2V^{\frac{1}{3}}\mu$. For the d.c. resistance $R_{dc} = \rho N^2\lambda/k_\omega W$ where λ is the mean length of turn, W the total cross-sectional area of the winding space, and k_ω the fraction occupied by the conductor. Here again for a core of fixed shape $W/\lambda = V^{\frac{1}{3}}$ and $R_{dc} = k_2N^2/V^{\frac{1}{3}}$ where k_2 is a constant. Eliminating N we have

$$R_{dc} = \frac{k_3L}{V^{\frac{1}{3}}\mu} \text{ and } \quad D_{dc} = \frac{k_4}{V^{\frac{1}{3}}\mu f}$$

The values of D for the eddy-current, hysteresis, and residual losses can be written down at once from the formula on page 237 of the October Editorial, and we have for the first four items

$$D = \frac{k_4}{V^{\frac{1}{3}}\mu f} + k_5\mu f + k_{10}\frac{\mu^2L^2I}{V^{\frac{1}{3}}} + k_7\mu$$

where I is the r.m.s. current and the k terms are constants. For given values of L, I and f the optimum permeability can be found by assuming a number of values of μ and plotting the calculated values of D, but this cumbersome procedure is not

necessary if some of the terms are known to be negligibly small, which is usually the case. In the article referred to it is shown that if only R_{dc} and R_r need be considered, then by differentiating with respect to μ it is found that

$$\mu_{opt} = \frac{k_4^{\frac{1}{3}}}{k_7^{\frac{1}{3}} V^{\frac{1}{3}} f^{\frac{1}{3}}} \text{ and } D_{opt} = 2\frac{k_4^{\frac{1}{3}} k_7^{\frac{1}{3}}}{V^{\frac{1}{3}} f^{\frac{1}{3}}}$$

μ_{opt} is found to be the value that makes $R_{dc} = R_r$.

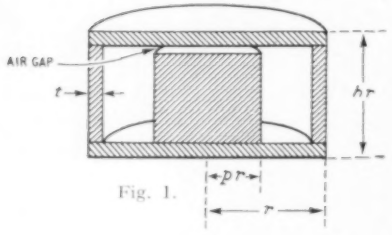

Fig. 1.

It is somewhat surprising to find that the assumption that the eddy current and hysteresis losses can be neglected in comparison with the residual magnetic losses is usually applicable. This is partly due to the high resistivity of the ferrites and partly to the fact that when used with transistors the power level is so low that the hysteresis loss becomes negligible. If used under other conditions such that R_{dc} and R_h predominate, then

$$\mu_{opt} = \sqrt[5]{\frac{4k_4^2}{9k_{10}^2 f^2 L I^2 V^{\frac{1}{3}}}} \text{ and}$$

$$D_{opt} = \left[\left(\frac{3}{2}\right)^{\frac{2}{3}} + \left(\frac{2}{3}\right)^{\frac{2}{3}}\right] \sqrt[5]{\frac{k_4^3 k_{10}^2 L I^2}{f^3 V^3}}$$

μ_{opt} is here the value that makes $R_{dc} = 1{\cdot}5R_h$.

The condition that only R_{dc} and R_e need be considered is usual with ordinary magnetic material but is of little importance with ferrites. When it is applicable

$$\mu_{opt} = \frac{k_4^{\frac{1}{3}}}{k_5^{\frac{1}{3}} V^{\frac{1}{3}} f} \text{ and } D_{opt} = \frac{2k_4^{\frac{1}{3}} k_5^{\frac{1}{3}}}{V^{\frac{1}{3}}}$$

As in the case of the residual losses, for μ_{opt}, $R_{dc} = R_e$.

If in these three cases D/D_{opt} is plotted against μ/μ_{opt} it is seen that for variations of μ/μ_{opt} between 0·8 and 1·2, D/D_{opt} does not vary more than 2 or 3%. It must be remembered that μ represents effective permeability, which can be adjusted by varying the air-gap. In all the above formulae the volume V of the core appears in the denominator showing that the losses can be reduced by increasing the size of the core, but there are limitations to this due to the two items which we have omitted, viz., R_c and R_s due to capacitance and a.c. losses in the wire. The skin effect and eddy-current losses can, of course, be reduced by finer stranding of the wire, and the authors mention a case in which the wire consisted of 810 insulated strands, but this involves a considerable increase in the d.c. resistance. The effects of distributed capacitance throughout the coil are very complex and as the author says "do not lend themselves to representation in practicable generalized formulas". Experiments were made and it was found that in the case of a very small specially-constructed inductor coil the distributed capacitance reduced Q from 330 to 300, while in a much larger inductor of standard construction the Q for d.c. resistance and core losses was reduced from 1,000 to 630 by replacing the solid wire by a stranded wire, which meant a reduction of the space factor k_w from 0·4 to 0·13, and the Q was further reduced from 630 to 550 due to the distributed capacitance.

Another question that arises is that of the best shape of the core. Assuming the type shown in Fig. 1 with the air-gap anywhere in the magnetic circuit, and the cross-section of the outer wall equal to that of the centre post, what should be the values of p and h to give the best results with a given overall volume? By expressing t, A, l, W, λ, and the coil volume in terms of r, p, and h, and inserting these in the above formulae, the dissipation factor D can be calculated for various assumed values of p and h, and can then be plotted against p for any given value of h. When hysteresis loss can be neglected the curves show a minimum D when p is about 0·45 and h is 1·2; when hysteresis loss is the predominant core loss these figures are 0·5 and 1·0, which differ so little from the former values that one may regard the optimum relative dimensions as independent of the nature of the core losses. Fig. 1 is drawn with the former values.

The adjustment of the effective permeability by varying the air-gap and by other methods is discussed in the article referred to. Instead of a plane gap it is suggested that a conical gap, as shown in Fig. 2, should be used to give a finer adjustment

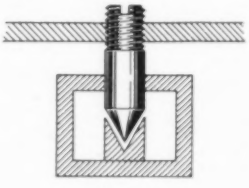
Fig. 2.

As an example of the improvement due to the use of ferrites it is stated that a standard type of inductor is only about a third of the size of the nearest equivalent coil with a permalloy core, but has over twice the Q value.

G. W. O. H.

OSCILLATOR CHARACTERISTIC EQUATION

Theory and Experimental Verification

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SUMMARY.—An equation representing the oscillation characteristics of a triode-oscillator circuit is theoretically developed, involving differential coefficients of the dynamic resistance of the triode. It is experimentally verified using a paralled-fed short-wave Hartley oscillator, and almost complete agreement is shown. Some interesting features of the oscillation characteristics are discussed and the constants of the triode and the tank circuit used are deduced therefrom.

Introduction

THE study of the oscillation characteristics[1,2] of an oscillator circuit is of great importance from the standpoint of its performance and the dynamic stability of the oscillations generated. When the oscillations are being set up and are gradually moving towards the steady state, the dynamic anode resistance r_a of the triode, and therefore the anode conductance g_a cannot be regarded as constant under the circumstances, as supposed in the elementary theory leading to a first-degree equation connecting the anode current with anode voltage for a given operating bias. The object of the present paper is to develop theoretically a second-degree equation representing oscillation characteristics involving the first differential coefficient of r_a and to show that this equation agrees with the experimentally obtained curves, except in a portion on the positive v_a side.

Equation of Oscillaton Characteristic

The anode current I_a of a triode in an oscillatory circuit may be expressed as some function of anode voltage V_a and grid bias V_g by the following relation, called the oscillation characteristic by Appleton[1,2]

$$I_a = \phi(V_a, V_g) \quad .. \quad .. \quad .. \quad (1)$$

where
$$\left.\begin{aligned} I_a &= I_0 + i_a \\ V_a &= V_0 + v_a \\ V_g &= V_{g0} + v_g \end{aligned}\right\} \quad .. \quad .. \quad (2)$$

I_0, V_0, V_{g0} representing the steady values and i_a, v_a, v_g the oscillatory components. To determine the above function ϕ, use may be made of Taylor's series expansion.[3] Thus we have the expression for the oscillatory component of anode current in a triode with a resistance load R_a in its anode circuit,

$$i_a = \frac{\mu}{r_a + R_a} \cdot v_g - \frac{\mu^2 r_a}{2(r_a + R_a)^3} \cdot \frac{\delta r_a}{\delta v_a} \cdot v_g^2 + \left[\frac{\mu^3 r_a}{6(r_a + R_a)^5} \left\{ (2r_a - R_a) \left(\frac{\delta r_a}{\delta v_a} \right)^2 - r_a (r_a + R_a) \frac{\delta^2 r_a}{\delta v_a^2} \right\} \cdot v_g^3 \right] \quad .. \quad (3)$$

where μ is amplification factor of the triode, treated as constant. This equation may now be applied to a triode functioning as an oscillator; e.g., with a tank circuit inserted between the anode and filament. Under these circumstances we have,

$$v_g = -\alpha v_a \quad .. \quad .. \quad .. \quad .. \quad (4)$$

α being a positive constant less than one, and v_a and v_g being respectively the oscillatory anode potential and the grid feedback voltage in anti-phase, as required to sustain the oscillations in the tank circuit. Also R_a is now the effective impedance of the tank circuit at resonance.

Treating r_a as variable and its first derivative as constant under the operative conditions, so that

$$\frac{\delta^2 r_a}{\delta v_a^2} = 0 \quad .. \quad .. \quad .. \quad .. \quad (5)$$

the coefficient of v_g^3 in equation (3) becomes

$$\frac{\mu^3 r_a}{6(r_a + R_a)^5} (2r_a - R_a) \left(\frac{\delta r_a}{\delta v_a} \right)^2$$

The ratio of this coefficient to that of v_g^2 in the same equation is

$$= \frac{\mu}{3(r_a + R_a)^2} (2r_a - R_a) \left(\frac{\delta r_a}{\delta v_a} \right)$$

and

$$= \frac{\mu}{12 r_a} \frac{\delta r_a}{\delta v_a}$$

when the impedance of the tank circuit is matched to the dynamic anode resistance of the triode. This ratio being inversely proportional to r_a, which is of the order of some thousands, we might neglect the third-degree and subsequent high-power terms in comparison.

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Retaining the terms up to second degree in the expansion (3) and adding the steady anode current,

$$I_0 = \phi\,(V_0, V_{g0}) \qquad .. \quad .. \quad .. \quad (6)$$

we have from (4)

$$I_a = I_0 - \frac{\mu\alpha}{(r_a + R_a)} \cdot v_a - \left\{\frac{\mu^2\alpha^2 r_a}{2(r_a + R_a)^3} \cdot \frac{\delta r_a}{\delta v_a}\right\} \cdot v_a^2 \qquad .. \quad (7)$$

For the impedance of the tank circuit at resonance,

$$R_a = \frac{4\pi^2 f^2 L_a^2}{R} \qquad .. \quad .. \quad .. \quad (8)$$

where L_a is the inductance of anode coil and R the series resistance in the tank circuit oscillating at frequency f. Substituting the above value of R_a in equation (7) we obtain finally the oscillation characteristic (i.e., the function ϕ explicitly in terms of v_a) as

$$I_a = I_0 - \left\{\frac{\mu\alpha}{\left(r_a + \frac{4\pi^2 f^2 L_a^2}{R}\right)}\right\} \cdot v_a - \left\{\frac{\mu^2\alpha^2 r_a}{2\left(r_a + \frac{4\pi^2 f^2 L_a^2}{R}\right)^3} \cdot \frac{\delta r_a}{\delta v_a}\right\} \cdot v_a^2 \qquad .. \quad .. \quad .. \quad (9)$$

From equation (9), it is clear that the shape of oscillation characteristic described by the function ϕ must approximate to a parabolic curve. The various constants occurring in its equation must necessarily depend on the operating conditions of the oscillator as defined by relations (4), (5), (6) and (8).

Experimental

To verify the validity of the above conclusion a parallel-fed Hartley circuit was used in the experimental work. The oscillation characteristics were determined by the method described by Appleton (loc. cit.) and also by another method devised by the present author.[4] Close agreement between the results obtained by these two methods for a Hartley circuit was previously reported.[4,5] In the present work the same Hartley oscillator was used but, to ensure greater accuracy, the mean of the results obtained by these two methods has been taken. Any serious grid loading of the oscillator was avoided by suitable choice of grid-blocking capacitor and grid-leak resistance, so as to ensure class A operation.

The constants of the tank circuit used were as follows:

Wavelength $\lambda = 80$ metres.
Coefficient of grid feedback voltage $\alpha = 0{\cdot}54$.
Inductance of the anode coil $L_a = 18{\cdot}58\ \mu$H.

Oscillation characteristics were determined for four different steady anode potentials, 60, 100, 140 and 200 V. These are graphically shown in Fig. 1.

Discussion of the Results

All the four oscillation characteristics shown graphically in Fig. 1 are largely of parabolic shape except the portions AB which lie on the positive v_a side, in spite of the tendency towards positive skewness. For the major portion BCD of each oscillation characteristic, which lies on the negative v_a side, a second-degree curve is fitted by the statistical method of least squares. The equations of the best fitting curves are as follows:

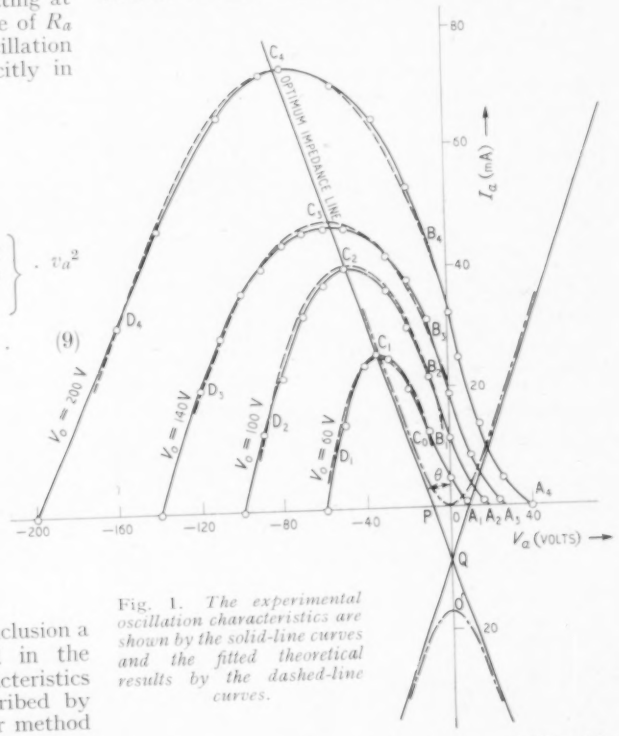

Fig. 1. *The experimental oscillation characteristics are shown by the solid-line curves and the fitted theoretical results by the dashed-line curves.*

for $V_0 = 60$ V,

$$I_a = 1{\cdot}1 - 1{\cdot}46\,v_a - 0{\cdot}02333\,v_a^2 \qquad .. \quad (10.1)$$

for $V_0 = 100$ V,

$$I_a = 11{\cdot}45 - 1{\cdot}28\,v_a - 0{\cdot}01398\,v_a^2 \qquad (10.2)$$

for $V_0 = 140$ V,

$$I_a = 22{\cdot}05 - 0{\cdot}864\, v_a - 0{\cdot}00731\, v_a^2 \quad (10.3)$$

for $V_0 = 200$ V,

$$I_a = 33{\cdot}8 - 1{\cdot}015\, v_a - 0{\cdot}00650\, v_a^2 \quad (10.4)$$

where I_a is expressed in milliamperes and v_a in volts.

These are shown as dotted curves over the corresponding experimental characteristics. The agreement between the experimental and the fitted second-degree curves, as defined by general equation (9), is extremely close in each case. In spite of uncertain factors like overheating of the valve, etc., the validity of equation (9) to represent the oscillation characteristic of an oscillator is thus established.

The equation (9) may be used to determine the value of $\delta r_a/\delta v_a$ for the triode under oscillating conditions. Let σ denote ratio of the coefficient of v_a^2 to the square of that of v_a in equations (10). Then using equation (9),

$$\sigma = \frac{r_a \times 10^{-3}}{2(r_a + 4\pi^2 f^2 L_a^2/R)} \cdot \frac{\delta r_a}{\delta v_a} \quad .. \quad (11.1)$$

and assuming the impedance of the tank circuit to be matched to the dynamic resistance of the triode, resulting in maximum power in the tank circuit,

$$\sigma = \frac{10^{-3}}{4} \cdot \frac{\delta r_a}{\delta v_a} \quad .. \quad .. \quad .. \quad (11.2)$$

The values of σ as determined from the first three of equations (10) are 0·011, 0·009 and 0·010 giving the mean value of $\delta r_a/\delta v_a = 40$ ohms per volt, within the range of voltage used on the anode. This value may be taken to represent the order of magnitude, since power in the tank circuit may not be quite maximum under the working conditions. Incidentally, it may be mentioned that the above values of σ obtained from equations (10), which represent the experimental results, justify the assumption as regards the constancy of $\delta r_a/\delta v_a$ and hence the relation (5).

Some Features of Oscillation Characteristics

We note from equation (9) that I_a passes through a maximum corresponding to a certain voltage v_a'. Differentiating (9) and equating to zero we have,

$$v_a' = -(r_a + 4\pi^2 f^2 L_a^2/R)^2/\mu x\, r_a \cdot \frac{\delta r_a}{\delta v_a} \quad (12.1)$$

and under the matched condition,

$$v_a' = -\frac{4r_a}{\mu x} \Big/ \frac{\delta r_a}{\delta v_a} \quad .. \quad .. \quad .. \quad (12.2)$$

Since $\delta r_a/\delta v_a$ is already evaluated and v_a' can be read off from the relevant experimental oscillation characteristic, the relations (12) may be used to find r_a under the given operating condition.

As seen from the figure, these maximum points for the four curves lie on a straight line $C_1 C_2 C_3 C_4$. This point is also verified from relations (10) which lead to the following equation for this line:

$$I_a = -7{\cdot}8 - 1{\cdot}014\, v_a \quad .. \quad (13.1)$$

giving the intercepts $p = \mathrm{OP} = 7{\cdot}7$ V and $q = \mathrm{OQ} = 7{\cdot}8$ mA. It may, however, be noted that for the oscillation characteristics corresponding to very small values of V_0, maxima cannot lie on this line but should fall on a smooth curve $\mathrm{OC_0C_1}$ passing through the origin O, since in the ultimate stage the smallest oscillation characteristic will vanish, merging in the point O. However, long before this stage is reached the oscillations will cease. Thus we might regard the locus of maxima as one continuous curve O C_0 C_1 C_2 C_3 C_4 having the line given by equation (13.1) as its asymptote. As a first approximation, this curve may be treated as a part of a hyperbola whose equation is found, with the help of (13.1), to be,

$$p^2 I_a^2 + 2p^2 q I_a = q^2 v_a^2$$

From this we get:

(i) for the initial bent portion $\mathrm{OC_0}$ of the curve, neglecting I_a^2 in comparison to $2q$,

$$I_a = \frac{q v_a^2}{2p^2} \quad .. \quad .. \quad .. \quad .. \quad (13.2)$$

(ii) for that part of the curve which is nearly straight,

$$I_a = -q - \frac{q}{p} v_a \left\{1 + \tfrac{1}{2} \frac{p^2}{v_a^2}\right\} \quad .. \quad (13.3)$$

where v_a is large compared with p. This finally degenerates into the straight line given by (13.1) for comparatively very large values of v_a. Relations almost identical with (13.2) and (13.3) have been empirically obtained recently by Singh.[6]

Thus the equations (13) completely describe in parts the locus of maximum of oscillation characteristics under various operating conditions. A physical interpretation may be given to the curve O $C_0C_1C_2C_3C_4$ by taking its gradient at a point as a measure of the dynamic stability of oscillations in the valve circuit operating under conditions represented by that point. The inverse of this gradient, which then is interpreted as the total impedance of the oscillator (the triode and the tank circuit taken together) is seen to diminish at first as the operating point moves up the curve, and then to become constant after reaching an oscillation characteristic of a certain value of V_0. This corresponds to the region where the straight portion of the curve begins. This stage, therefore, represents the optimum total impedance which has the smallest value, and thus the beginning of the dynamic stability of oscillations of a given frequency. From this point of view, it is unprofitable to operate the oscillator at a higher anode

potential than this optimum value since dynamic stability of oscillations does not improve further. The asymptote QC_4, therefore, may be looked upon as the optimum impedance line of the oscillator for the frequency f. The optimum impedance is given by $\tan \theta$, where θ is the angle between the asymptote and the axis of anode current.

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VOLTAGE-REFERENCE NODE

Its Transformations in Nodal Analysis

By Jacob Shekel

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SUMMARY.—This paper describes a general method of analysing linear networks by nodal analysis, without specifying the node to which all voltages are referred. When the reference node is specified, or when the voltage-reference node is changed, the admittance matrix of the network undergoes certain transformations. The method is applied to valve circuits, where it enables a simpler approach to earthed-grid or earthed-anode stages, or circuits where no electrode is earthed.

Introduction

WHEN a network is analysed by nodal equations,[1,2,3] it is necessary to choose one of its nodes as a reference node, and the voltages of all other nodes are measured (or defined) relative to this special node. A network with n nodes is analysed by $n-1$ independent equations, and may be represented by a square admittance matrix of order $(n-1) \times (n-1)$. This paper discusses the transformation of this matrix due to a change of the reference node, and also proposes a method of treating networks without specifying the reference node at all.

The outlined method has a special application in analysing valve circuits. The usual procedure is to represent a valve by a passive network and internal voltage or current sources. When matrix algebra is applied,[4,5] the treatment follows that of two-terminal-pair networks; the earthed-cathode, earthed-grid and earthed-anode stages need special derivation of the matrix components. In the proposed method, an n-electrode valve is treated as an n-terminal network, and the derived general transformations are applied to represent change of earthing point.

"Indefinite" Admittance Matrix

Consider a network with n nodes. Let $\boldsymbol{I}$ be a column matrix, whose ith component I_i is the external current entering the ith node. Let I_{ik} denote the current from node i to node k through the $i-k$ branch . ($I_{ik} = -I_{ki}$, $I_{ii} = 0$.)

Applying Kirchhoff's Current Rule to the ith node,

$$I_i + \sum_{k=1}^{n} I_{ki} = 0$$

Summing the equations for all n nodes,

$$\sum_i I_i + \sum_i \sum_k I_{ki} = 0 \; (i, k = 1, 2 \ldots n)$$

The elements under the double summation cancel out in pairs, leaving

$$\sum_i I_i = 0 \qquad \ldots \qquad \ldots \qquad \ldots \qquad (1)$$

Let $\boldsymbol{V}$ be the column matrix whose ith component V_i is the voltage of the ith node, relative to an arbitrary reference voltage. (The latter may be the voltage of any node, a combination of some of them, or any other voltage.) Since only voltage differences determine the network currents, the latter must be invariant to the addition of an arbitrary voltage V_0 to all the V_i.

In linear networks, $\boldsymbol{I}$ and $\boldsymbol{V}$ are related linearly:

$$\boldsymbol{I} = \boldsymbol{Y}\,\boldsymbol{V} \qquad \ldots \qquad \ldots \qquad \ldots \qquad (2)$$

where $\boldsymbol{Y}$ is a square matrix of order $n \times n$.

Put all components of $\boldsymbol{V}$ equal to zero, except V_j, then

$$I_i = Y_{ij} V_j$$

$$\sum_i I_i = \sum_i (Y_{ij} V_j) = \left(\sum_i Y_{ij} \right) V_j$$

and, by (1)

$$\left(\sum_i Y_{ij} \right) V_j = 0.$$

As this is to be true for any V_j,

$$\sum_i Y_{ij} = 0 \qquad \ldots \qquad \ldots \qquad \ldots \qquad (3)$$

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Suppose now that an arbitrary voltage V_0 is added to all the components of $\boldsymbol{V}$. The current into the ith node will then be

$$I_i = \sum_k Y_{ik}(V_k + V_0) = \sum_k Y_{ik}V_k + \left(\sum_k Y_{ik}\right)V_0,$$

and if the currents are to be invariant to the addition of any V_0,

$$\sum_k Y_{ik} = 0 \quad .. \quad .. \quad .. \quad (4)$$

Equations (3) and (4) show that the sum of any row and column of the matrix $\boldsymbol{Y}$ is zero. This matrix, which relates the currents with voltages that are referred to an undefined reference point, will be termed 'the indefinite admittance matrix'. It evidently is a singular matrix and, as the solution of network problems usually calls for an inversion of the matrix, some transformation will be necessary.

General Transformation

A general method for impedance matrix transformations is described by Kron.[6] His results, when applied to admittance matrices, are as follows[7]:

Let the 'old' voltages (before the transformation) $\boldsymbol{V}$ be described as a linear combination of 'new' voltages $\boldsymbol{V}'$,

$$\boldsymbol{V} = \boldsymbol{A}\boldsymbol{V}' \quad .. \quad .. \quad .. \quad (5)$$

($\boldsymbol{A}$ may be a rectangular matrix, so that $\boldsymbol{V}$ and $\boldsymbol{V}'$ do not necessarily have the same number of components.)

To keep the form for power ($P = \boldsymbol{I}_t\boldsymbol{V}$) invariant, 'new' currents $\boldsymbol{I}'$ must be defined by

$$\boldsymbol{I}' = \boldsymbol{A}_t\boldsymbol{I} \quad .. \quad .. \quad .. \quad (6)$$

where $\boldsymbol{A}_t$ is the transpose of $\boldsymbol{A}$; i.e., the matrix $\boldsymbol{A}$ with its rows and columns interchanged.

Finally, to keep the admittance equation (2) invariant, the 'new' admittance matrix must be defined by

$$\boldsymbol{Y}' = \boldsymbol{A}_t\boldsymbol{Y}\boldsymbol{A} \quad .. \quad .. \quad .. \quad (7)$$

This general transformation will now be applied to particular cases pertaining to change of reference node.

Specifying the Voltage-Reference Node

Suppose node 4 of a four-node network is chosen as the voltage-reference node. (The discussion may easily be extended to the general case of an n-node network, and it seems unnecessary to present the derivation in general terms.) The voltages are then defined by

$$V'_i = V_i - V_4 \ (i = 1, 2, 3) \quad .. \quad (8)$$

For convenience, V_4 is taken equal to zero, and (8) becomes

$$\left.\begin{aligned} V_i &= V'_i \ (i = 1, 2, 3) \\ V_4 &= 0 \end{aligned}\right\} \quad .. \quad (9)$$

In matrix notation, (9) is the same as

$$\boldsymbol{V} = \boldsymbol{A}\boldsymbol{V}', \text{ where } \boldsymbol{A} = \begin{Vmatrix} 1 & 0 & 0 \\ 0 & 1 & 0 \\ 0 & 0 & 1 \\ 0 & 0 & 0 \end{Vmatrix} \quad .. \quad (10)$$

The corresponding transformation of the admittance matrix is

$$\boldsymbol{Y}' = \boldsymbol{A}_t\boldsymbol{Y}\boldsymbol{A} = \begin{Vmatrix} 1 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 \\ 0 & 0 & 1 & 0 \end{Vmatrix} \times \begin{Vmatrix} Y_{11} & Y_{12} & Y_{13} & Y_{14} \\ Y_{21} & Y_{22} & Y_{23} & Y_{24} \\ Y_{31} & Y_{32} & Y_{33} & Y_{34} \\ Y_{41} & Y_{42} & Y_{43} & Y_{44} \end{Vmatrix} \times \begin{Vmatrix} 1 & 0 & 0 \\ 0 & 1 & 0 \\ 0 & 0 & 1 \\ 0 & 0 & 0 \end{Vmatrix}$$

$$\boldsymbol{Y}' = \begin{Vmatrix} Y_{11} & Y_{12} & Y_{13} \\ Y_{21} & Y_{22} & Y_{23} \\ Y_{31} & Y_{32} & Y_{33} \end{Vmatrix} \quad .. \quad .. \quad .. \quad (11)$$

This proves the following theorem:

When any node is specified as the voltage-reference node, the corresponding row and column are to be omitted from the indefinite admittance matrix.

Changing the Voltage-Reference Node

Once the admittance matrix is given in a definite form (with the reference node specified), it may be desirable to choose a different node for reference. This may be carried out in two steps: first, the matrix is brought to the indefinite form, and then the new reference node is specified. The first step is treated below. (The second step was described in the preceding section.)

Let us again take a 4-node network, described by a 3×3 matrix, with node 4 as a reference node; and let us define the new voltages $\boldsymbol{V}'$ by

$$V_i = V'_i - V'_4 \ (i = 1, 2, 3) \quad .. \quad (12)$$

In matrix notation, this is

$$\boldsymbol{V} = \boldsymbol{A}\,\boldsymbol{V}', \text{ where } \boldsymbol{A} = \begin{Vmatrix} 1 & 0 & 0 & -1 \\ 0 & 1 & 0 & -1 \\ 0 & 0 & 1 & -1 \end{Vmatrix} \quad .. \quad (13)$$

The transformation of the admittance matrix is then

$$\boldsymbol{Y}' = \boldsymbol{A}_t\boldsymbol{Y}\boldsymbol{A} = \begin{Vmatrix} 1 & 0 & 0 \\ 0 & 1 & 0 \\ 0 & 0 & 1 \\ -1 & -1 & -1 \end{Vmatrix} \times \begin{Vmatrix} Y_{11} & Y_{12} & Y_{13} \\ Y_{21} & Y_{22} & Y_{23} \\ Y_{31} & Y_{32} & Y_{33} \end{Vmatrix} \times \begin{Vmatrix} 1 & 0 & 0 & -1 \\ 0 & 1 & 0 & -1 \\ 0 & 0 & 1 & -1 \end{Vmatrix}$$

$$\mathbf{Y}' = \begin{Vmatrix} Y_{11} & Y_{12} & Y_{13} & -Y_{11}-Y_{12}-Y_{13} \\ Y_{21} & Y_{22} & Y_{23} & -Y_{21}-Y_{22}-Y_{23} \\ Y_{31} & Y_{32} & Y_{33} & -Y_{31}-Y_{32}-Y_{33} \\ -Y_{11}-Y_{21}-Y_{31} & -Y_{12}-Y_{22}-Y_{32} & -Y_{13}-Y_{23}-Y_{33} & \sum_{i=1}^{3}\sum_{j=1}^{3} Y_{ij} \end{Vmatrix} \quad (14)$$

We have thus proved the theorem:

To bring the definite admittance matrix to indefinite form, one row and one column are to be added, whose elements are such that they complete the sum of each row and column to zero.

Short-Circuiting Two Nodes

A similar type of transformation, though not dealing directly with the reference node, may be mentioned here.

Suppose that in a 4-node network, described by a 4×4 admittance matrix, nodes 3 and 4 are short-circuited together. This may be represented by introducing 'new' voltages $\mathbf{V}'$

$$\left.\begin{aligned} V_i &= V'_i \\ V_4 &= V'_3 \end{aligned}\right\} \quad (i = 1, 2, 3) \quad .. \qquad .. \qquad (15)$$

In matrix notation

$$\mathbf{V} = \mathbf{A}\mathbf{V}', \text{ with } \mathbf{A} = \begin{Vmatrix} 1 & 0 & 0 \\ 0 & 1 & 0 \\ 0 & 0 & 1 \\ 0 & 0 & 1 \end{Vmatrix} \quad .. \qquad .. \qquad (16)$$

giving a new admittance matrix

$$\mathbf{Y}' = \mathbf{A}_t\,\mathbf{Y}\,\mathbf{A} = \begin{Vmatrix} 1 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 \\ 0 & 0 & 1 & 1 \end{Vmatrix} \times \begin{Vmatrix} Y_{11} & Y_{12} & Y_{13} & Y_{14} \\ Y_{21} & Y_{22} & Y_{23} & Y_{24} \\ Y_{31} & Y_{32} & Y_{33} & Y_{34} \\ Y_{41} & Y_{42} & Y_{43} & Y_{44} \end{Vmatrix} \times \begin{Vmatrix} 1 & 0 & 0 \\ 0 & 1 & 0 \\ 0 & 0 & 1 \\ 0 & 0 & 1 \end{Vmatrix}$$

$$\mathbf{Y}' = \begin{Vmatrix} Y_{11} & Y_{12} & Y_{13}+Y_{14} \\ Y_{21} & Y_{22} & Y_{23}+Y_{24} \\ Y_{31}+Y_{41} & Y_{32}+Y_{42} & Y_{33}+Y_{34}+Y_{43}+Y_{44} \end{Vmatrix} \quad .. \quad (17)$$

Theorem: When any two nodes are short-circuited together, the corresponding rows of $\mathbf{Y}$ are added together to form one row, and the corresponding columns are added together to form one column.

Had node 4 been an 'isolated node' [see node 4 in Fig. 3(a)] shorting it to any other node would have had no effect on the network, so that we may formulate the following corollaries:

1. A row and column of zeros correspond to an isolated node.
2. Any n-node network may be treated as an m-node one ($m > n$), by adding $m - n$ rows and columns of zeroes to the admittance matrix.

Application to Valves

There are two general methods of approach to linear valve circuits (small-signal approximation). In one method, the valve is represented by a passive network, to which internal voltage or current sources have been added. The second method,[2,8] which is more readily applicable to matrix methods, represents the valve by an unsymmetric admittance (or impedance) matrix.

The general $(n + 1)$ electrode valve (not counting the heater), with the cathode as the voltage-reference node, is described by an $n \times n$ admittance matrix[8]

$$Y_{ij} = \frac{\partial I_i}{\partial V_j} \quad .. \qquad .. \qquad .. \qquad .. \qquad (18)$$

In the simplest case, a triode operating in class A, let the grid and anode be numbered 1 and 2 respectively, then

$$\mathbf{Y} = \begin{Vmatrix} 0 & 0 \\ g_m & g_a \end{Vmatrix} \quad .. \qquad .. \qquad .. \qquad (19)$$

(g_m is the mutual conductance, and g_a is the reciprocal of the anode resistance r_a).

The indefinite matrix is obtained by completing each row and column to zero, as in Equation (14):

$$\mathbf{Y}' = \begin{Vmatrix} 0 & 0 & 0 \\ g_m & g_a & -g_m - g_a \\ -g_m & -g_a & g_m + g_a \end{Vmatrix} \quad .. \quad (20)$$

This matrix is the starting point for calculating earthed-grid or earthed-anode stages, by omitting the row and column corresponding to the earthed electrode.

When the triode is connected as a diode, the internal resistance depends on the connection:

1. If the grid is connected to the cathode, this corresponds to earthing terminal 1 of (19). Crossing out the first row and column leaves a conductance g_a.
2. When the grid is connected to the anode, row 1 is added to row 2, and column 1 to column 2, giving a conductance $g_a + g_m$.

(The values of g_a and g_m in both cases are different, due to different steady voltages on the grid.)

Illustrative Example

To illustrate the procedure formulated above, let us analyse a network for d.c. voltage stabilization, shown (in essentials) in Fig. 1.

Fig. 2 shows the equivalent network for voltage changes (disregarding constant voltage differences), and differs from Fig. 1 in the following points:

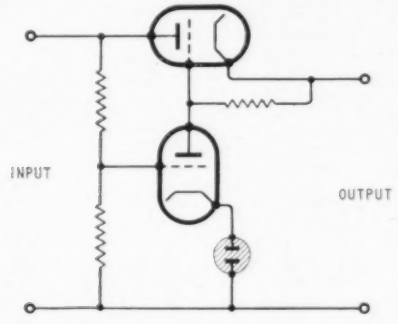

Fig. 1. *Voltage stabilizer.*

1. The cathode of V_2 is earthed, as the gas-filled diode keeps it at a constant voltage level.
2. The grid of V_2 is assumed to draw no current, therefore the voltage divider may be replaced by the relation $V_4 = kV_1$, where k is a constant. This will result in a transformation of the admittance matrix.

The admittance matrix will be constructed in three steps.

1. The matrices of the constituent parallel networks (Fig. 3) are written down. $\boldsymbol{Y}_b$ and $\boldsymbol{Y}_c$ are derived from (20), omitting the row and column of the earthed node (in $\boldsymbol{Y}_c$), and changing rows and columns to conform with the numbering of the nodes. Rows and columns of zeroes are added to all matrices, corresponding to the isolated nodes.

$$\boldsymbol{Y}_a = \begin{Vmatrix} 0 & 0 & 0 & 0 \\ 0 & G & -G & 0 \\ 0 & -G & G & 0 \\ 0 & 0 & 0 & 0 \end{Vmatrix}$$

$$\boldsymbol{Y}_b = \begin{Vmatrix} g'_a & g'_m & -g'_m & -g'_a & 0 \\ 0 & 0 & 0 & 0 & 0 \\ -g'_a & -g'_m & g'_m & +g'_a & 0 \\ 0 & 0 & 0 & 0 & 0 \end{Vmatrix}$$

$$\boldsymbol{Y}_c = \begin{Vmatrix} 0 & 0 & 0 & 0 \\ 0 & g''_a & 0 & g''_m \\ 0 & 0 & 0 & 0 \\ 0 & 0 & 0 & 0 \end{Vmatrix}$$

Primed and double-primed g terms refer to V_1 and V_2 respectively.

2. All three matrices are added together to form the admittance matrix of the network.

$$\boldsymbol{Y} = \begin{Vmatrix} g'_a & g'_m & -g'_a - g'_m & 0 \\ 0 & G + g''_a & -G & g''_m \\ -g'_a & -G - g'_m & G + g'_a + g'_m & 0 \\ 0 & 0 & 0 & 0 \end{Vmatrix}$$

Both steps, of course, could have been taken together, writing down $\boldsymbol{Y}$ at once by inspection of Fig. 2.

3. The four voltages $\boldsymbol{V}$ are expressed by three new voltages $\boldsymbol{V}'$,

$V_1 = V'_1$, $V_2 = V'_2$, $V_3 = V'_3$, $V_4 = kV'_1$,

corresponding to a transformation matrix

$$\boldsymbol{A} = \begin{Vmatrix} 1 & 0 & 0 \\ 0 & 1 & 0 \\ 0 & 0 & 1 \\ k & 0 & 0 \end{Vmatrix}$$

Transforming $\boldsymbol{Y}$ to $\boldsymbol{Y}' = \boldsymbol{A}_t\boldsymbol{Y}\boldsymbol{A}$ has the effect of multiplying the fourth row and column by k and adding them to the first row and column respectively:

$$\boldsymbol{Y}' = \begin{Vmatrix} g'_a & g'_m & -g'_a - g'_m \\ kg''_m & G + g''_a & -G \\ -g'_a & -G - g'_m & G + g'_a + g'_m \end{Vmatrix}$$

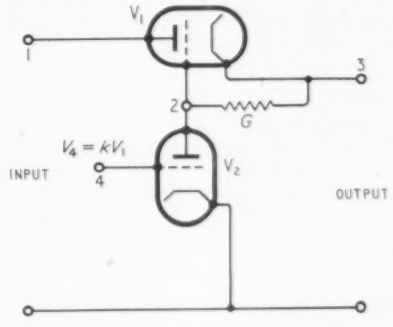

Fig. 2. *Equivalent circuit for voltage changes.*

We will now find the voltage-division factor k that is necessary for satisfactory operation.

There should be no change in the output voltage, even when the input voltage does change. This calls for a transfer impedance $Z_{31} = 0$. But as Z_{31} is the co-factor of Y'_{13} divided by the determinant of $\boldsymbol{Y}'$, it is sufficient if the co-factor itself be zero:

$$\begin{vmatrix} kg''_m & G + g''_a \\ -g'_a & -G - g'_m \end{vmatrix} = 0$$

$$k = \frac{g'_a\ (G + g''_a)}{g''_m\ (G + g'_m)}$$

On the other hand, it is desirable that the output voltage be independent of the output current, which calls for an output impedance $Z_{33} = 0$ (Thévenin's Theorem). The numerator of Z_{33} is the co-factor of Y'_{33}, so that

$$\begin{vmatrix} g'a & g'm \\ kg''m & G + g''a \end{vmatrix} = 0$$

$$k = \frac{g'a\,(G + g''a)}{g''m\,g'm}$$

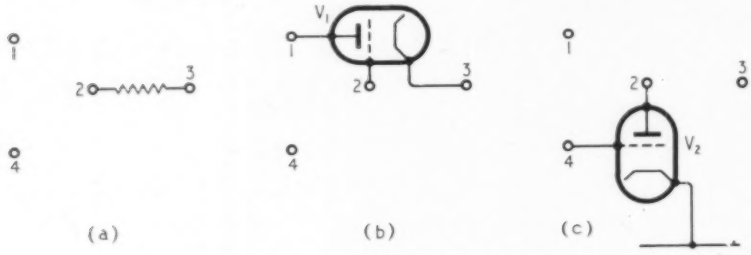

Fig. 3. *Constituent networks of the circuit shown in Fig. 2.*

Both values of k are incompatible, showing that, with the circuit of Fig. 1, both results cannot be achieved with one setting of k; but, the lower the value of G, the better may both results be approximated with a single setting.

Conclusion

A general method of nodal analysis without specifying a voltage reference node has been described. The main results are the following:

1. A network whose reference node is not specified may be represented by an indefinite admittance matrix. In this matrix the sum of every row and column is zero.

2. An indefinite matrix may be brought to definite form by omitting the row and column corresponding to the reference node.

3. A definite matrix is brought to indefinite form by adding a row and a column, so as to make the sum of every row and column equal to zero.

4. Changing the reference node is carried out by combining the two previous steps.

The method is applicable to valve circuits, where it enables easy manipulation of the valve constants, no matter which one of its electrodes, if any, is earthed. A similar treatment may be applied to transistor circuits,[9] or to any other multi-terminal network element.

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[6] G. Kron, "Tensor Analysis of Networks," Wiley and Sons, 1939.
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[8] Ibid, Chap. XV, "Multi-electrode Tube Circuits," p. 385.
[9] J. Shekel, "Matrix Analysis of Transistor Circuits," *Proc. Inst. Radio Engrs*, November 1952. Vol. 40, p. 1394.

INDEXES

The Index for 1953 to the editorial pages of *Wireless Engineer* will be included in the March 1954 issue, in which there will also be the Index to Abstracts and References published during 1953, and a list of journals scanned for abstracting, with their publishers' addresses.

The March 1954 issue, which will include the normal editorial pages, as well as the Indexes, will be priced at 6s. The Index pages will be detachable for binding with the 1953 volume.

H.F. DIRECTION FINDING

Comparison of C.W. and Pulsed Transmissions

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SUMMARY.—The causes of errors in high-frequency direction finding are briefly discussed and numerical estimates are made of the error-components when pulsed and continuous-wave transmissions are utilized at frequencies in the region of 8 Mc/s. Particular attention is devoted to the performance of Adcock cathode-ray direction finders in the absence of a ground ray; given adequate signal strength, pulse-operation will normally be at least as accurate as c.w. operation. For a well-maintained instrument of this type, installed on a good site and manned by a first-class operator, it is concluded that under the most favourable circumstances for pulse-operation (i.e., when a direct first-order echo via the E or E_s layer can be used) the expected ratio of pulse variance to c.w. variance is about one-third for single snap bearings: for bearings averaged over a few minutes, the expected improvement from pulse operation is rather less. These tentative estimates refer to transmission distances from about 400 to 2,000 km. It is noted that ionospheric conditions do not permit the whole of this range to be covered at all times by the modes of propagation mentioned above.

Introduction

THE purpose of the paper is briefly to consider whether any improvement in bearing accuracy would be likely to result from the use of pulse-modulated transmissions instead of continuous waves for practical direction-finding purposes in the high-frequency band. The study was stimulated by a question accepted by the International Radio Consultative Committee of the International Telecommunication Union at the Geneva meeting in 1951 (Question 61).[1] Special interest arises at frequencies around 8 Mc/s, as survival craft wishing to communicate with the maritime mobile service in connection with search and rescue operations have, by international agreement, been allocated the frequency of 8·364 Mc/s.

Except within a limited radius of a high-frequency transmitter, the signals from it are received solely by way of the ionosphere and errors in bearing-measurement are introduced which are absent when only the ground ray is in use. In addition to polarization errors, which are of particular concern with loop direction finders, other types of error arise in both loop and Adcock instruments which depend upon the complexity of the ionospheric signal. It is natural, therefore, to consider whether any advantage in accuracy would be gained by using pulsed transmissions which, within certain limits, enable a single 'ray' to be selected from the composite signal. It is interesting to note that the possibilities of pulsed transmissions as a means of improving the accuracy of h.f. direction finding were realized as long ago as 1932 by Eckersley and Smith in a British Patent;[2] the subject is also referred to in the Annual Report for the same year of the Radio Research Board of the Department of Scientific and Industrial Research[3] and in a review of cathode-ray oscillography published in 1933.[4] Since that date much use has been made of pulse transmissions in directional studies of ionospheric waves for research purposes but considerations of bandwidth and interference, and uncertainty as to the countervailing improvement in accuracy to be achieved have been among the factors which have militated against the use of pulses in practical direction-finding work. The question of accuracy is discussed here; an analysis is made of the various sources of directional errors and a tentative estimate is given of the improvement in direction-finding accuracy at high frequencies to be expected by using a single ray. Because of its limited range, at any rate in ground–ground working overland, attention is mainly directed to the case when the ground-wave is absent.

MS accepted by the Editor, April 1953

2. Discussion of Causes of D.F. Errors

Since there is not perfect horizontal stratification of ionic density, the reflecting surface in the ionosphere is neither horizontal nor smooth. Moreover, the structure of the surface is not stationary but wave-like motions occur;[5,6] these waves show a wide variety of characteristics—from undulations of considerable vertical amplitude and some hundreds of kilometres in wavelength, to much smaller disturbances having dimensions comparable with the radio wavelength. The tilt of the wave-surfaces from the horizontal is generally small, amounting to only one or two degrees.[7]

Dealing with reception of a single 'ray' reflected at the ionosphere, we may for simplicity consider a first-order ordinary ray; that is, one which has undergone a single reflection and from which the extraordinary magneto-ionic component is absent due to absorption. Observations made with a direction finder which is free from errors depending on the polarization of the radiation accord-

ingly show comparatively slow bearing deviations (lateral deviation) corresponding to reflection at a large-scale sloping surface on which are superposed rapid fluctuations due to scattering from the fine structure. It is convenient to note here that the lateral deviation introduced by a given angle of tilt at the reflecting surface increases with the angle of elevation of the ray arriving at the receiver; that is, bearing error increases with the height of reflection, with order of reflection, and with decreasing distance of the transmitter. An additional effect which causes trouble when observing a ray of higher order than the first is the scattering occurring at the ground reflection point or points. The scattering due to the ionospheric fine-structure, which has already been mentioned, results in radiation being received over a finite range of angle; scattering at the ground reflection point increases this cone-angle and consequently the magnitude of the rapid bearing fluctuations.

Suppose now, as usually happens in practice, that conditions allow of more than one mode of propagation between transmitter and receiver. For example, there may be two or more orders of reflections from the F layer, a single-order reflection in which the extraordinary ray is not negligible in intensity compared with the ordinary, or reflections from both the E and F layers. If continuous-wave signals are in use these various components will not be resolved in time and consequently the direction finder intercepts at any instant two or more narrow cones of radiation, neither of which, in general, will be centred in the great-circle plane between transmitter and receiver. Neglecting the effect of the 'coning' for the moment, consider two plane waves of comparable amplitude arriving at slightly different azimuths. The observed bearing will depend on the angular separation of the rays and on their relative phases and amplitudes. The largest deviations occur when the waves are in antiphase and for a limited range of phase angle the deviation may amount to many times the angular separation of the rays; for the remainder of the phase-range the deviation is in the opposite sense and of the order of half the magnitude of the azimuthal separation of the rays. These wave-interference errors, due to the simultaneous reception of more than one ray, result in a considerable scatter of the bearings; they contribute to the rapid component of bearing fluctuations. To the causes of error discussed above must be added those remaining after local calibration has provided corrections for certain instrumental effects and for imperfections of the site in the immediate neighbourhood of the direction finder. These residual errors include observational errors, polarization errors, and errors arising from radiation scattered from features many wavelengths distant from the direction finder.

3. Comparison of Pulse and C.W. Direction Finding

It has been shown above that errors in high-frequency direction finding using continuous waves can arise from the following causes:

(i) Observational (human element).
(ii) Instrumental (including polarization errors, and those arising from the site close to the direction finder).
(iii) Distant site errors.
(iv) Lateral deviation in the ionosphere.
(v) Wave interference due to 'coning' of each ray and between different propagational modes.

In the following paragraphs the relative importance of these various sources of error will be assessed quantitatively. A statistical approach will be used, a *variance* (or mean-square) figure being attached to each class of error for c.w. and pulse transmissions. This approach is convenient for, if the assumption is made that the various error-classes are independent of each other, the overall variance is the sum of the component class-variances. The complexity and variability of propagation conditions render it difficult to assign specific values under certain of the headings; however, this has been attempted by reference to the published literature which, it may be noted, deals mainly with reception conditions in England. In making estimates, two distinct methods of observation have been borne in mind: the first, in which a single instantaneous bearing is taken; and the second, in which the mean of, say, 10 such snap bearings taken over a period of 5 minutes is utilized. It is assumed that a well-maintained fixed U-type Adcock cathode-ray direction finder is installed on a good site[8] and is used by a skilled operator. Attention has been concentrated on distances in excess of about 400 km; it will, however, be possible to arrive at certain limited conclusions relating to much shorter distances when the ground-ray is present, and also when loop direction finders are in use. There is no point in considering cases where the ionospheric signal reaches the receiver by other than normal modes since, when these break down, direction finding using c.w. or pulse transmissions is without value.

(i) *Observational Errors*

Except in very adverse circumstances the standard deviation of a single bearing (for a skilled operator) will be less than a degree[9] with c.w. operation; if the mean of a number of bearings is taken there will be a corresponding reduction. There will be some increase in observational error when the bearing is fluctuating rapidly. It will, therefore, be assumed that variances of 1 deg^2 and zero refer to a single observation and multiple observations respectively in c.w. operation.

With pulse operation the figure of 1 deg^2 quoted above may be somewhat reduced; the variance of the mean of a number of observations will still approximate to zero. It should be noted that the increased bandwidth necessary may increase interference from stations on neighbouring frequencies (and, therefore, observational error) unless adequate transmitter power is available.

(ii) *Instrumental Errors*

It is assumed that the direction-finding station is carefully maintained and frequently calibrated by means of a local transmitter.

The residual errors in this class then arise from (a) changes in the instrument and its immediate surroundings occurring since the most recent calibration, (b) defects in calibration technique due to the proximity of the transmitter, (c) uncertainty of the appropriate instrumental correction to apply when the dominant mode of propagation (and, therefore, elevation-angle) is not known, and (d) polarization errors.

The major contribution is probably from (a) on a site of high conductivity.

With c.w. operation it is estimated that these various factors together introduce a variance of 1 or 2 deg^2 for single bearings on a site of high conductivity. Little advantage is gained by averaging a number of bearings. For poor conductivity sites, higher variances for a single bearing will be observed but averaging the results will tend to eliminate this contribution from the polarization-error component.

Pulse operation can be expected to give little improvement from ray-selection. If, however, the site is of poor conductivity some advantage will be gained by using a low-angle ray due to the reduction in polarization error; however, the improvement will be slight when averaging of bearings over a few minutes is permissible.

(iii) *Distant Site Errors*

This effect is estimated to contribute, with c.w. operation, a variance of about 1 deg^2 for single and averaged bearings on a good site in the band 5–10 Mc/s;[10,11] the effect is more important at the lower frequencies of the h.f. band.

No improvement would be expected from pulse operation.

(iv) *Lateral Deviation*

A figure of about 2 deg^2 may be taken for this effect with c.w. operation for day-time working and about 4 deg^2 for night-time.[12] This figure is assumed to apply equally to single snap observations and to the mean of a number of observations over a 5-minute period since lateral deviation represents a slowly varying component of variance.

In the case of pulse operation, if the range of transmission and ionospheric conditions is such that propagation by a first-order E or E_s reflection is possible then this ray would naturally be chosen. In these circumstances it is estimated[7] that the variance contribution would be approximately 0·5 deg^2; but if no such echo is available the higher angle of elevation of the other rays will result in a negligible improvement compared with c.w. working. As in the case of c.w. operation the estimated variances refer both to single and averaged snap bearings.

(v) *Wave Interference*

With c.w. operation, the errors due to wave interference represent the major contribution to the rapidly-varying errors. A representative figure for the variance of a single snap bearing is taken to be about 4 deg^2, although this shows wide variations depending on the complexity of the received radiation. For snap bearings averaged over a period of 5 minutes the variance contribution will be approximately 0·5 deg^2.

When a single mode is selected by pulse operation there is left mainly the variance contribution due to 'coning.' For first order E or F echoes the variance will be small both for single and averaged bearings.[7] An upper limit of 0·5 deg^2 is assigned to this contribution.

4. Conclusions

(a) The Table on the next page summarizes the information given in the previous section relating to the components of variance at a frequency in the region of 8 Mc/s for a well-maintained U-Adcock cathode-ray direction finder installed on a good site and used by a first-class operator. In considering the Table it must be remembered that the various estimates given are of necessity rather rough and that they refer to ionospheric reception at distances greater than 400 km. The Table—as indeed the paper in general—relates to conditions commonly noted in the United Kingdom. It cannot, unfortunately, be assumed that all the data are representative of conditions in other parts of the world and throughout the sunspot cycle.

(b) The Table shows that under normal conditions, pulse-operation with adequate transmitter power appears capable of providing bearings at least as accurate as those obtainable with c.w. working. At ranges from the transmitter such that ground and ionospheric waves are comparable in intensity, errors on c.w. working will be high and might, if pulsed transmissions are used instead, be reduced by ground-ray selection to a variance of about 2–4 deg^2. This figure applies to an Adcock system, but an improvement due to ground-ray selection would be expected also for loop direction finders which might well be in use in ships or aircraft during rescue operations. When

no ground-ray is present and an Adcock direction finder is in use, the Table shows that in the most favourable circumstances for pulse-working (i.e., *when a direct first-order echo via the E or E_s layer of adequate strength is receivable*) the expected ratio of pulse to c.w. variances is about one-third for single snap bearings corresponding to a ratio of about 0·6 in standard deviation: for snap bearings averaged over a few minutes a slightly smaller improvement is indicated. It must again be emphasized that these figures are tentative and that any deterioration of the observer's ability, and of the site and equipment from the high standards indicated in the Table will reduce any potential advantages of pulse operation relative to c.w.

Cause of error	Variance (deg²)			
	Continuous Wave		Pulse	
	Single snap bearing	Mean of 10 snap bearings in 5 minutes	Single snap bearing	Mean of 10 snap bearings in 5 minutes
(i) Observational	1	0	0–1	0
(ii) Instrumental	1–2	1	1–2	1
(iii) Distant site errors	1	1	1	1
(iv) Lateral deviation	2 (day) 4 (night)	2 (day) 4 (night)	0·5 (1E or 1E_s in use, otherwise, as for C.W.)	0·5 (1E or 1E_s in use, otherwise, as for C.W.)
(v) Wave interference	4	0·5	0–0·5	0–0·5
Totals	9–10 (day) 11–12 (night)	4·5 (day) 6·5 (night)	2·5–5 (1E or 1E_s in use)	2·5–3 (1E or 1E_s in use)

(c) It is outside the scope of this paper to discuss whether the range of circumstances in which the improvement is likely to be obtained is worth the increased complexity of apparatus, higher demands on operating skill, and the cost involved in money and spectrum space. It may be mentioned, however, that in the United Kingdom on about 8 Mc/s during day-time, first order E or E_s reflections are generally receivable in the range 400–2,000 km. At night the low-angle first-order echoes are from E_s and the smallest range of transmission available by this mode will vary seasonally and from night to night: although the upper limit of 2,000 km may be achieved throughout the year, given adequate transmitter power, the median minimum range may vary between 1,400 km in winter and 600 km in summer.

Acknowledgment

The paper is published by permission of Marconi's Wireless Telegraph Co., Ltd., and the Director of Radio Research of the Department of Scientific and Industrial Research.

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VALVE AND RECEIVER NOISE MEASUREMENT AT V.H.F.

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SUMMARY.—Techniques for the measurement of noise factor, valve noise and other parameters required in the investigation of receiver noise performance are described. Sources of error, and the precautions required to minimize them, are discussed. New equipment giving improved experimental accuracy is described.

Experimental results on noise generators and measurements of valve noise are included.

1. Introduction

METHODS of noise measurement have been described in a number of earlier papers,[1,2,3] and the techniques to be described in this paper are very similar. However, it is thought that a record of methods which have been used at T.R.E. over a number of years, during which time there have been numerous investigations of the accuracy, will be of value to other workers, in view of the growth of interest in the subject in late years.

General definitions will be found in a recent paper.[4]

2. Methods of Measurement

2.1. *Noise Factor*

Absolute Methods

A CV172 tungsten-filament diode has been used as the standard noise source. This valve gives true saturated noise from about 1,000 c/s to 100 Mc/s. An alternative type, the CV2171 on a B7G base, is now available.

The correct conditions of measurement of noise factor are that the receiver should be operated at an ambient temperature of 17°C (290°K = T_0) with a dummy aerial, or source, at the same temperature. In practice this is not always easy to achieve, and some correction must be applied. It is not possible to allow for this correction very precisely and when absolute accuracy is required it is desirable that the temperature should be not very different from the standard. The noise output from the receiver may be expressed thus:

Noise output = Receiver noise + $kTB.M$

$= V + kTB.M$

where: k = Boltzmann's constant

T = Temperature of source, degrees Kelvin

B = Noise bandwidth of the receiver

M = Power gain.

The noise factor is measured by finding the increase in available noise from the source which is required to double the noise output. Hence, using a saturated noise diode feeding a source resistance R_A, if the diode current required is I_d

$$V + kTBM = \frac{2e\, I_d\, B.R_A M}{4}$$

where e = electron charge.

Assuming that the receiver noise is independent of ambient temperature, the noise factor N at the standard temperature T_0 is:

$$N = 1 + \frac{V}{kT_0BM}$$

$$\text{Hence} \quad N = 1 + \frac{2eI_dR_A}{4kT_0} - \frac{T}{T_0} \quad \ldots \quad (1)$$

And for $T_0 = 290°\text{K}$, $\dfrac{2e}{4kT_0} = 0{\cdot}0200$

for I_d in mA and R_A in ohms.

In practice, the dependence of receiver noise on temperature is not known. Therefore, for cross-check measurements, when it is not possible to work at a temperature close to T_0, it is better to quote the temperature of measurement and give the result as

$$N \text{ (at temperature } T) = \frac{2eI_dR_A}{4kT_0} \times \frac{T_0}{T} \quad \ldots \quad (2)$$

For operational purposes, the measurement should always be made with the source at temperature T_0, and the receiver at normal operating temperature.

The receiver noise output must be measured before the second detector, and it is necessary to be able to measure a given increase in this noise output (usually to double the mean-square voltage).

Four methods have been used for this measurement:

(a) In theoretical investigation, special valve tests, etc., a test amplifier, with sufficient gain to swamp the noise from succeeding stages, is connected to a special amplifier which is terminated by a square-law anode-bend detector.

The special amplifier is built into a rack and

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includes a display for indicating interference.

With this unit measurement of noise factor within a frequency band of 120 kc/s (effectively the single-frequency noise factor) can be made over a range of frequencies of 20 to 90 Mc/s and also at 12 Mc/s.

It consists of a frequency changer (which can be switched as a low-gain 12-Mc/s amplifier), five stages of amplification at 12 Mc/s, and square-law valve voltmeter. All stages except the last are relatively wideband, and the last stage contains a filter circuit. A choice of two filter-circuit bandwidths is provided, one about 120-kc/s wide and the other about 500-kc/s wide with a flat response over 250 kc/s. Gain control is applied to five stages and a wide range of control is obtained with negligible change in bandwidth or linearity.

(b) For measurement of full-band noise factor a test amplifier with sufficient gain to swamp the noise from succeeding stages is connected to a succeeding amplifier of appropriate bandwidth and this includes a diode detector in its output.

The output of the head amplifier and input of the main amplifier are brought out through matching pads to coaxial connectors. By insertion of a pad of attenuation 2 (3 db) and by increasing the noise level of the source till the original output-meter reading is restored, the noise factor of the head amplifier can be measured. *The pad must give no change of bandwidth when inserted.*

(c) On a production receiver with diode detector it has been found fairly satisfactory to use a meter in the diode circuit, operating between output levels of 2·5 and 3·5 V (i.e., $\sqrt{2}:1$). This method assumes that the diode in its linear detection region reads r.m.s. volts of noise.

(d) If the gain control over a range of about 6 db gives negligible change in bandwidth, then an ingenious method due to Waltz[5] can be used. This eliminates errors in detector law at the expense of considerably increased errors due to any drifts in gain and meter readings. The gain is set to give a convenient output-meter reading and noise, corresponding to a noise-diode current I_1, is added to increase the output to a second convenient reading. The gain is then reduced to restore the output-meter reading to the first value, and the noise-diode current increased to a value I_2 to bring the output to the second reading. Then the equivalent value of I_d in equations (1) and (2) is

$$I_d = \frac{I_1^2}{I_2 - 2I_1}$$

For methods (a), (b) and (c), the noise output ratio can be calibrated using a noise-diode generator, and a correction to the term containing I_d in equation (1) is $\frac{1}{A-1}$ where A is the true ratio of mean-square output voltage.

It is only necessary to arrange that the input noise is very much greater than the receiver noise and the law can be checked directly in terms of noise-diode current. With (a), since the check is of small deviations from true square-law only, it is fairly satisfactory if the receiver noise is as much as 10 % of the total and the calibration is made by backing-off the output reading with no additional noise at the input. Alternatively, the square-law detector can be checked very satisfactorily by using two c.w. sources of slightly different frequencies both feeding the receiver input and with adequate padding to ensure that the input from either one to the receiver is independent of the other. Then the two levels may be set equal to each other by switching on each oscillator alternately, and with both oscillators on together the output reading should be doubled. It is necessary to check over a bigger d.c. range than that used because the peaks of noise reach higher values than a sine-wave.

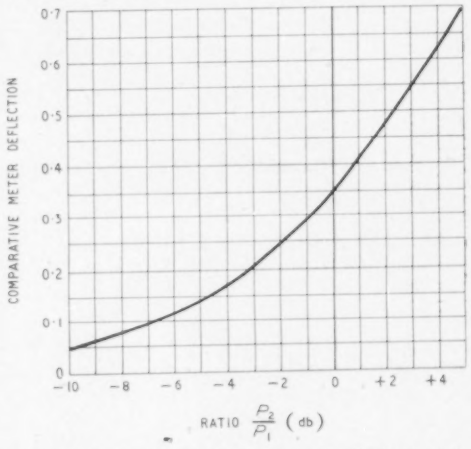


Fig. 1. *Direct-reading noise-factor unit. Scale shape law; P_1 = receiver noise power, P_2 = peak injected noise power.*

Direct-Reading Instrument

For production testing of valves and amplifier units, and in development work, a direct-reading unit is extremely valuable. This enables circuits to be adjusted to optimum conditions easily and accurately. The method is not absolute except on differences, but the output can be calibrated by using a transfer standard. The noise source is modulated by a square-wave to ensure saturated current during the 'on' period. The modulation frequency should preferably be greater than any supply frequency components and sufficiently high to suit the design of a simple band-pass filter. The mean level of noise at the demodulator is maintained by long-time-constant a.g.c. circuits, and the rectified output is passed through a modula-

tion-frequency band-pass filter to a further rectifier feeding a d.c. meter. The depth of modulation of the noise at the input to the demodulator, and hence the output meter reading, is inversely proportional to receiver noise.

By suitable choice of levels, the meter calibration gives a scale which is approximately linear in decibels (Fig. 1). A discrimination of 1% in noise factor is readily obtained by careful filtering to avoid flicker due to noise components.

If the modulation frequency is low, considerable filtering may be required to give freedom from meter flicker with a consequent long-time-constant of meter. This is due to the change in spectrum of noise after rectification giving an increase in noise density at low frequencies.

Balanced System

In measuring the noise factor of an i.f. amplifier which is to be used with balanced mixers, the noise generator should be connected to either mixer and an identical impedance connected to the other. Reversing the connections gives a check on the symmetry of the unit. The true noise factor, assuming the asymmetry is not due to any asymmetry in available power transfer (e.g., different losses at i.f. in the mixers) is then given by:

$$N = \frac{N_1 N_2}{N_1 + N_2} \approx \frac{N_1 + N_2}{4}$$

where N_1 is the noise factor calculated from the formula (1) for the first connection neglecting the second mixer, and N_2 that calculated for reversed connections. This can be shown as follows:

$$N = \frac{N_2}{1 + \frac{R_1}{R_2}} = \frac{N_1}{1 + \frac{R_2}{R_1}}$$

where R_1 and R_2 are the transformed source resistances at the grid.

Hence: $\frac{N_1}{N_2} = \frac{R_2}{R_1}$ and the rest follows.

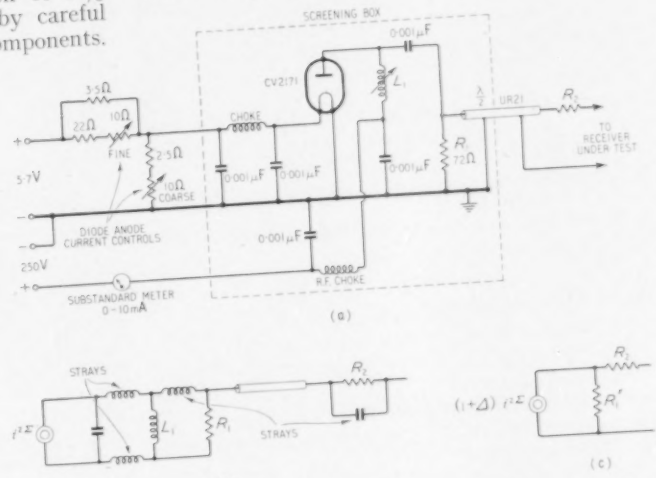


Fig. 2. *Noise diode circuit (a), and equivalent circuits (b) including strays and (c) for calculation.*

2.2. *Partial Noise Factors*

For experimental investigations it is sometimes desirable to measure the noise factor of a separate stage of an amplifier, usually the first. The pad attenuator method [2.1 (b)] gives a true measurement of the noise factor of stages preceding the pad, but it is not very practical when the output has to be padded up at the input to the attenuator (to ensure that the attenuator is giving true readings), since this padding may affect the noise factor. However, if the attenuator pad is calibrated under true working conditions, results are accurate and it is not then necessary to arrange for an accurate match at the input to the attenuator provided there is no bandwidth change.

A more convenient method is to switch off the stage or stages to be measured and replace these by a passive network of the same output impedance. The ratio Y' of the two noise outputs then gives:

$$N_1 = N\left(1 - \frac{1}{Y'}\right) + \frac{1}{M_1}$$

where N_1 is the noise factor of the early stage or stages,

N is the overall noise factor,

and M_1 is the available power gain of the early stages.

In practice, $1/M_1$ is usually negligible, and

$$N_1 = N\left(1 - \frac{1}{Y'}\right) \quad .. \quad .. \quad (3)$$

If M_1 is known, then N_2 may be calculated from

$$N_2 = (N - N_1)M_1 + 1 \quad .. \quad .. \quad (4)$$
$$= NM_1/Y'$$

2.3. *Noise-Diode Generator*

The circuit used is shown in Fig. 2. Although there are certain precautions to be taken to avoid

errors, this arrangement has been found to give satisfactory results and gives much greater flexibility for various applications than other methods which might be adopted for special experimental work.

For measurements of noise factor, the series resistor can be changed to obtain variation of noise factor with source resistance. This series resistance is usually mounted in the input circuit of the receiver on experimental amplifiers, but it can also be arranged as a mount of the same physical dimensions as a mixer crystal. It is then possible to measure noise factor of a unit which is connected to a crystal mixer, and very properly include the effect of any i.f. losses in the mixer. It is advisable to blow air on this resistor to keep its temperature close to ambient when making actual measurements.

If the temperatures of the two resistors are T_1 and T_2, the noise factor is given by:

$$N_1 = 0{\cdot}02 I_d \frac{R_1^2}{R_1 + R_2}\left(\frac{T_1 - T_0}{T_0}\frac{R_1}{R_1 + R_2} + \frac{T_2 - T_0}{T_0}\frac{R_2}{R_1 + R_2}\right)$$

where I_d is in milliamperes, R in ohms.

Possible errors with this measurement arise from the following causes:

(a) Noise diode not perfectly saturated or impedance not high compared with R_1.

(b) Transit time and lead inductance effects in the noise diode.

(c) Differences in bandwidth for thermal noise and diode noise at the input.

(d) Differences between d.c. and r.f. impedances of R_1 and R_2 or errors in measurement of resistance values. Shunting effect of circuit losses in parallel with R_1. Lead inductances in series with R_1 and capacitance in parallel with R_2.

(e) Errors due to the cable and plugs between R_1 and R_2.

To minimize all these errors the following precautions are necessary:—

(a) Operate with an anode voltage of at least 150. At this voltage the anode resistance of the CV172 is of the order of 10^5 ohms with a current of 10 mA.

(b) The 'lead inductance' effects occur chiefly in the thin filament and cannot be eliminated in a valve of this type. The correction factor is approximately of the form*:

$$i^{2\Sigma} = 2eI_d df\left(\frac{1}{1 - Af^2}\right)^2$$

or, $$i^{2\Sigma} \approx 2eI_d df(1 + 2Af^2)$$

Experimental tests have been made on very few samples, but these showed that A was variable between samples, and that a correction of about 1% may be required at 30 Mc/s and 10% at 100 Mc/s. This effect is strictly due to the distributed nature of the diode, which is approximately equivalent to an open-circuited line. No tests have been made with anode earthed, and the output and load impedance connected one to the centre and the other to the common ends of the heater, thereby simulating a transmission line diode.

The transit time is negligible up to 100 Mc/s with 150 V on the anode.

(c) With $R_1 = 72$ ohms, the frequency response of the diode circuit is such that the attenuation is less than 1% at $\pm$ 5 Mc/s. The error is therefore negligible for receiver bandwidths less than 10 Mc/s.

(d) The error in R_1 is mainly due to the coil losses in parallel, and this error can be allowed for. The resistor type used for R_2 is an R.M.A.16 for values of R_2 less than 500 ohms and an I.R.C.-type MPM† for values of R_2 between 500 and 2,000 ohms. Experiments at 45 Mc/s showed the shunting capacitance to have negligible effect and it is believed that the r.f. resistance is equal to the d.c. value to better than 1% with short leads. If higher source impedances are required in measurement, it is necessary to use a transformer. If the coil L_1 is connected at the diode anode terminal, and the circuit is resonant at these terminals the lead inductance in series with R_1 has negligible effect on the available power at the end of the cable. If there is appreciable lead inductance, then adjustment of L_1 to give maximum noise output will give error in the same way as shown in (b) above. However, although correct adjustment of L_1 gives a known result for the available power from the cable, it is then necessary to know the series resistance R_1' of the transformed impedance to be used in the term expressing the attenuation due to R_2, and the true available power becomes:

$$\left(2eI_d df\frac{R_1}{4}\frac{R_1'}{(R_1' + R_2)}\right)$$

(e) A low-loss close tolerance cable such as UR21 should be used. The error in noise factor due to cable and plug mismatches will be in proportion to the transformation ratio of R_1 (or Z_1) through these, when $R_2 \gg R_1$ and will be zero when $R_2 = 0$. A half-wavelength cable is worth while if measurements are all at the same frequency, and it is not convenient to use a very short cable. The attenuation will be about 1% at 50 Mc/s.

Even when precautions (a) to (e) are taken additional errors may arise due to experimental

* The symbol $x^{2\Sigma}$ is used for mean-square value of x.

† This is a resistor originated by the International Resistor Co. of U.S.A., and is designed for low end-to-end capacitance.

inaccuracies in reading the receiver output meter. The low-frequency noise output of a square-law detector is inversely proportional to the bandwidth before rectification, for a given total noise input, and there might be appreciable flicker indicated with bandwidths less than 2 Mc/s. This can be reduced by using large time constants in the meter circuits or a well-damped meter movement. A meter of sub-standard accuracy should be used for measurement of noise-diode anode current.

When extreme accuracy is required in measurement of noise factor variations with source impedance, the numerous experimental cross-checks can be reduced if the diode generator impedance is kept constant and the transformation to the valve input varied directly. This is quite practical on a special experimental unit.

Similarly, the experimental corrections are reduced if a lower resistance is used in the diode generator and this is transformed, with negligible loss of available power, into the required equivalent source impedance. Unfortunately, the bandwidth of a quarter-wave transformer is hardly adequate for all requirements, and the loss is excessive with simple lumped-circuit transformers.

In these experiments it is usually more important to obtain accuracy in measurement of noise factor than in knowledge of source impedance, and the calculated transformation of the source impedance (from the turns ratio of a tapped transformer) has mainly been used. In experimental investigations of noise-factor theory particular pains must be taken to measure the impedance at the valve grid.

2.4. *Shot Noise and Partition Noise*

It is convenient to use the same equipment for measurement of shot noise in valves. The arrangement is shown in Fig. 3. A unit has been used in which the valveholders for test valves have been modified so that the anode or screen-grid pin can be plugged into the same socket whatever the valve connections. All other connections are decoupled to ground. The diode comparator circuit includes means for adjustment of the load impedance, so that the impedances at the two output sockets are equal. A head amplifier of low-noise factor is used with a coaxial input which can be connected to either output.

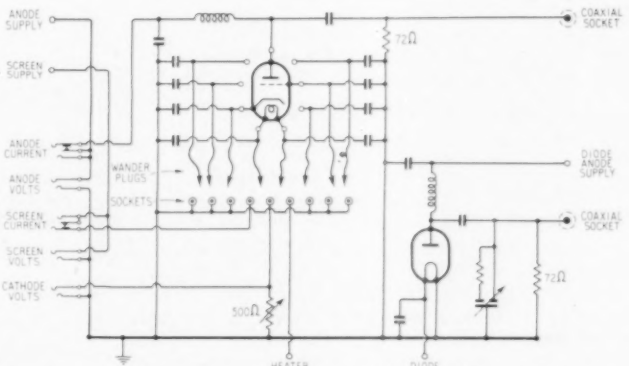


Fig. 3. *Arrangement for shot-noise measurements. The test holder connections are paralleled for B7G, BBA, etc., valve bases.*

This unit is arranged to have an optimum noise factor with a source impedance of 72 ohms, and is used in conjunction with the special noise receiver having a square-law output meter.

If the noise output is the same with the plug in either socket,

then $I_a F^2 = I_d$

where I_a = valve anode current

I_d = noise-diode anode current.

The errors in this measurement are as follows:—

(a) Lead inductance effects. These have been discussed for the noise diode. For the test valve they will be much smaller with modern type valves as they arise mainly from the actual leads and not the filamentary-cathode effect, as in the noise diode.

(b) Finite slope resistance r_a of the valve. A correction is necessary because this is not a generator of thermal noise as well as shot noise, so that

$$I_a F^2 = I_d\left(1 + \frac{1}{0{\cdot}020 I_d r_a}\right)$$

I_d is in mA and r_a in ohms.

For tetrodes or pentodes, a measurement of the noise in the anode and screen, gives a measure of

$$i_a^{2\Sigma} = F_0^2 2eI_a + \frac{(1 - F_0^2) I_{g2} 2eI_a}{I_a + I_{g2}}$$

$$i_{g2}^{2\Sigma} = F_0^2 2eI_{g2} + \frac{(1 - F_0^2) I_a 2eI_{g2}}{I_a + I_{g2}}$$

from which the separate contributions of shot and partition noise may be calculated. As a check the total shot noise may be measured separately by measuring the noise with an r.f. short between anode and screen.

2.5. *Noise Temperature Ratio*

In measurements of noise temperature ratio by direct methods, it is necessary to make a comparison with available noise from a resistance of the same value at standard temperature T_0. It is

also necessary to know the noise factor of the measuring unit. To simplify the measurement it is convenient if the result can be made less dependent on the exact equalization of the two resistances.

The actual comparison then is:

$$\frac{(N_1 + t_1 - 1)M_1B_1}{N_2M_2B_2}$$

The change of N with source can be minimized by choosing the correct mean source resistance at the valve grid, and B can be made as independent of source as we choose by making the bandwidth narrow at some late stage of the amplifier.

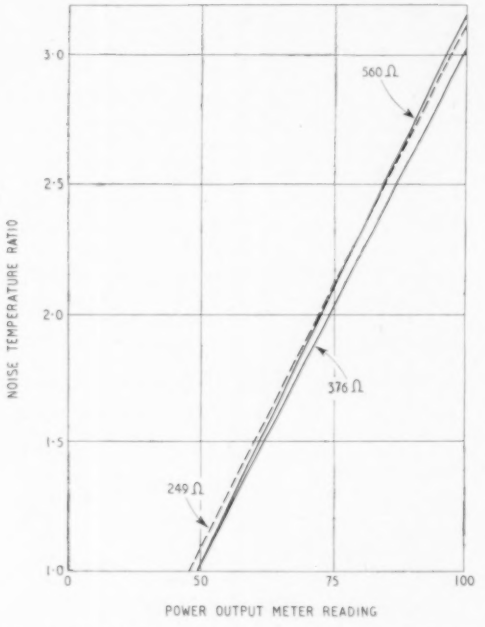


Fig. 4. *Calibration curves for measurement of noise temperature ratio.*

By making the input resistance of the valve equal to the source resistance giving optimum noise factor, the whole error is minimized, and this can be done using cathode-lead inductance. This gives the desired result with only a small increase in noise factor and consequently only small loss in discrimination, for the contribution from t_1 must be observed against a background of $(N_1 - 1)$.

Fig. 4 shows a typical calibration curve of a unit designed on this basis. The error in t for changes of source resistance is given by:

$$\delta t = (t - 1)x + N_2(x - y)$$

where x is the fractional change in gain, and y is the fractional change in noise factor; i.e.,

$$M_2 = M_1(1 + x) \text{ and } N_1 = N_2(1 + y)$$

By introduction of a variable, which can be tuned to maximum output on each measurement, it is possible to use the unit for a 10 : 1 range of resistance values.

This type of unit is also useful for measurements of radio-frequency noise when used with a 'matched' crystal mixer, as it minimizes the effects of changes in radio-frequency matching (or rather, mismatching).

2.6. *Conductance*

For investigations of noise factor it is often necessary to know the mutual conductance of the test valve and the conductance of various circuit elements.

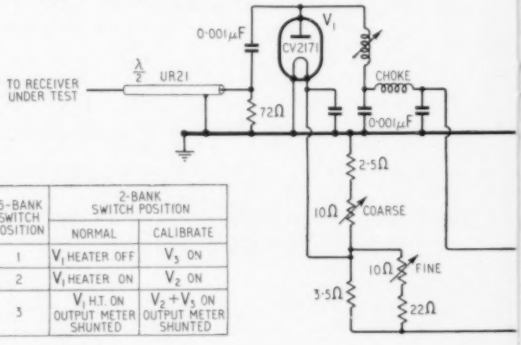
Mutual conductance has been measured at 750 c/s. The balancing resistors of the bridge have an accuracy of 1 part in 10^4 and balance discrimination is about 1 part in 10^3.

Conductances in the range 10 to 1,000 micromhos have been measured using a Q-meter method, with a unit specially made for the purpose. Calibration of this unit has been made with I.R.C.-type MPM resistors.

For equalization of admittances of the order of 10,000 micromhos an r.f. bridge has been used which gives a balance discrimination of 10 micromhos, or better, and which is limited by the stability of the adjustable components in the admittances being compared.

3. Description of Modern Equipment

New equipment has been constructed recently by Mr. C. T. Hodgson of R.R.E. and incorporates a number of improvements in stability and calibration methods. The measuring receiver can be switched over for use as an automatic gain-controlled amplifier and then becomes part of a direct-reading noise-factor meter.



5-BANK SWITCH POSITION	2-BANK SWITCH POSITION	
	NORMAL	CALIBRATE
1	V_1 HEATER OFF	V_3 ON
2	V_1 HEATER ON	V_2 ON
3	V_1 H.T. ON OUTPUT METER SHUNTED	$V_2 + V_3$ ON OUTPUT METER SHUNTED

Fig. 5 (and right). *Noise diode and meter shunt calibrator circuit diagram.*

The mains supplies are fully stabilized using a CV449 as reference for the h.t. and a CV172 as reference for the heater supplies. All valves are 'wired-in' except for the first-stage in the experimental head-amplifier in which a particular valve type is to be tested. These precautions have made a considerable improvement in stability, which is particularly noticeable when using the noise source in these days of unstable mains supplies. Regulation is superior to battery supplies (except with careful maintenance of the batteries). Other helpful features include the use of a double-triode for the square-law detector and 'backing-off' valve, overload protection for the output meter, and 'Helipot' controls to give wide range with fine control. In use, the zero of the output meter and gain of the amplifier often remain constant within setting accuracy over periods of hours.

The calibration of the output meter is accurate to about one part in 500. This accuracy is achieved by using a special unit for calibration, in which two CV172 valves are arranged to give an exact doubling of noise. The background noise level in this calibration is reduced to insignificant proportions, and the discrimination of changes on the output meter is better than 1 in 10^3. The two noise diodes are connected in parallel to a common-grid triode stage. The switching sequence is as follows:

(a) One diode on, the other biased beyond cut-off.
(b) The second diode on, the first cut-off.
(c) Both diodes on.

The diode on in (b) position has fixed heater supply and diode (a) has a fine control to adjust to the same anode current. A shunt across the output meter is switched in synchronism with the switching from (b) to (c) The procedure then is:

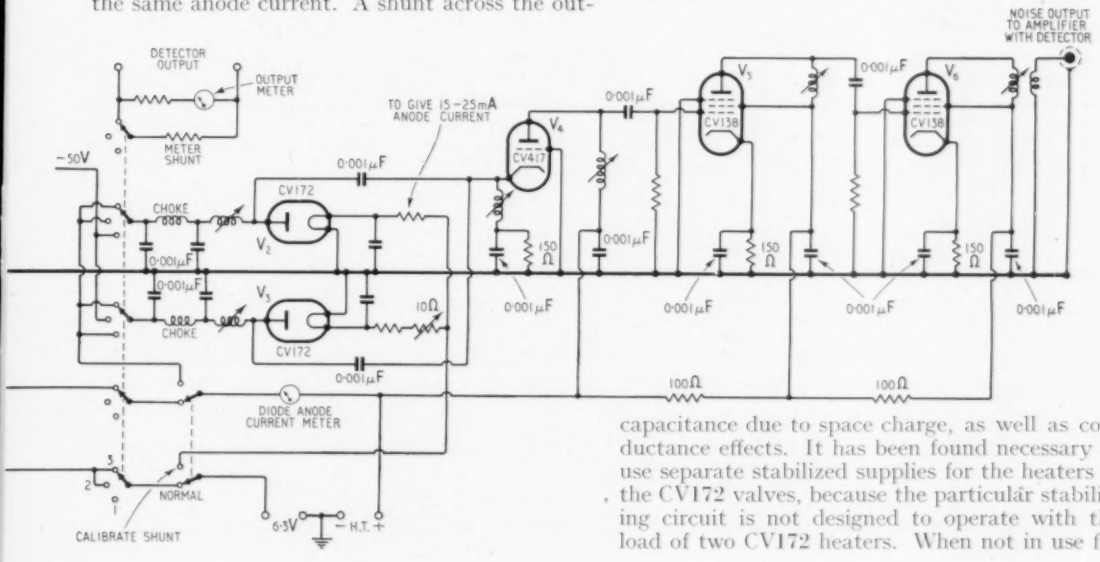
(i) Set calibration switch to (b) and adjust gain of main unit to give suitable output-meter reading. Switch off diode and check residual noise and 'back-off' if necessary. (By careful design, the residual noise can be made negligibly small.)
(ii) By switching from (b) to (a) and back, adjust both diodes to give exactly the same output.
(iii) By switching from (b) to (c) and back, adjust output-meter shunt to give exactly the same meter reading in both positions.

This method of calibration is preferable to using a 3-db pad and actually gives such a high accuracy as to be useful for the absolute calibration of attenuators.

Once the meter shunt is adjusted, the shunt connection is transferred to an identical switch in the noise-generator unit. This shunt is now switched in synchronism with the h.t. supply to the noise generator, so that the measurement of noise factor entails switching between two positions and adjusting the diode-generator heater current for no change in output-meter reading.

A Weston Type S82 meter 0–200 μA full-scale is used for the output, giving excellent discrimination. A similar type of meter is used for measuring the noise-generator diode current.

The double-diode unit uses a common-grid amplifier to minimize the effect of the finite anode conductance of a CV172 at a current of 15 mA and anode voltage 250. The 'off' valve must be biased back to eliminate changes in



capacitance due to space charge, as well as conductance effects. It has been found necessary to use separate stabilized supplies for the heaters of the CV172 valves, because the particular stabilizing circuit is not designed to operate with the load of two CV172 heaters. When not in use for

calibration, this unit supplies heater and anode voltages for test amplifier units. The circuit diagram is given in Fig. 5.

The test-amplifier units used in experiments on particular valve types are designed to give sufficient gain to swamp the noise of the main amplifier. The bias supplies to the first stage can be adjusted externally—either as cathode resistor or grid voltage, and a switch is fitted in the cathode return for use when neutralizing the grid–anode capacitance. The anode current of the test valve can be monitored separately. The input circuit contains a small tuning capacitor on the grid for measurements of capacitance-detuning effects, as well as enabling a high-Q low-loss coil to be used. The input coil is wound with a range of tapping points for changes of source impedance at the grid.

With the direct-reading test set it is possible to measure the effect of circuit changes, and changes of valve operating conditions, with ease and precision.

Outline circuit diagrams of the equipment are given in Fig. 6.

4. Experimental Results

4.1. *Diode Noise Generator*

Fig. 7 shows experimental results for the 'saturated' slope resistance of a CV172 at different operating currents and voltages. Figs. 8 and 9 show anode-current control characteristics for the CV2171 and CV172 (different presentations).

The effect of lead inductance (or rather the distributed nature of the diode) has been measured on two samples in terms of the transformation of anode–cathode conductance. If G_a is the total

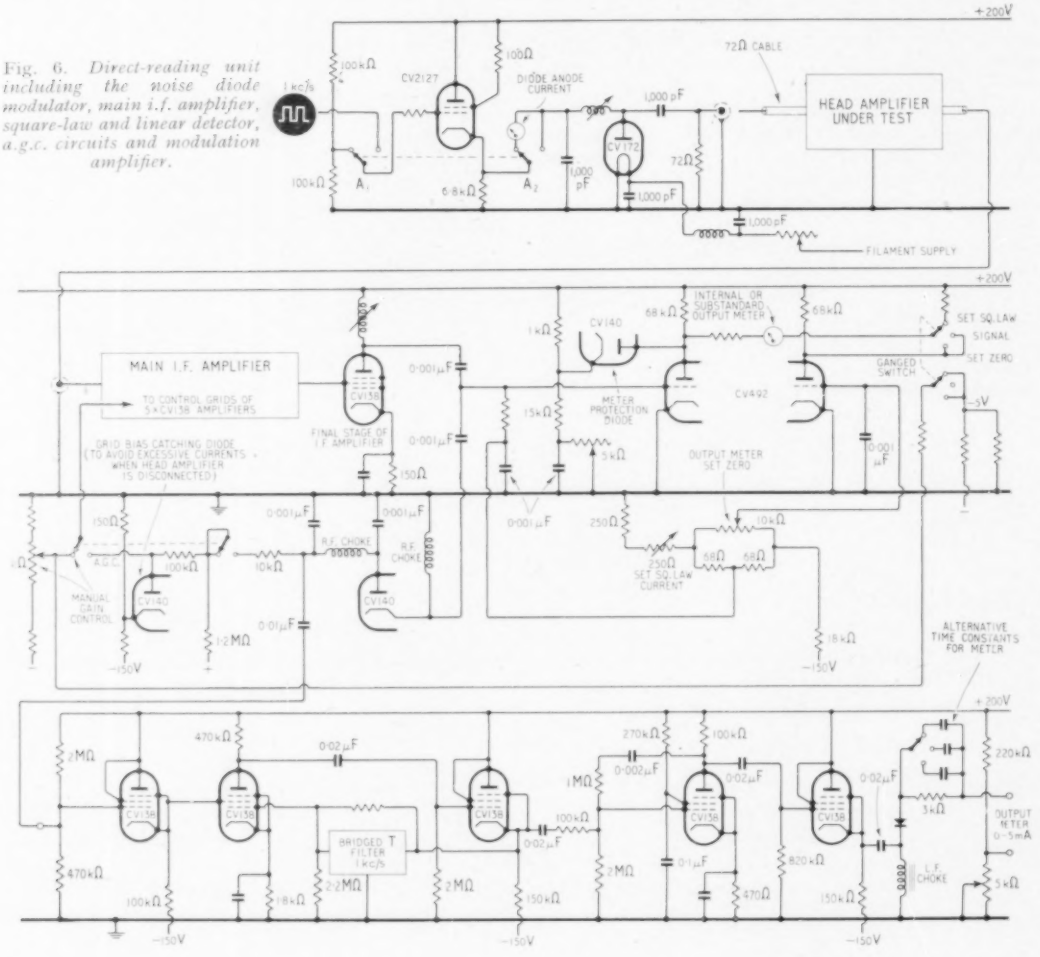


Fig. 6. *Direct-reading unit including the noise diode modulator, main i.f. amplifier, square-law and linear detector, a.g.c. circuits and modulation amplifier.*

conductance between anode and cathode, then the value G_a' measured at the output terminals is transformed in exactly the same way as the mean-square current generator, namely

$$G_a' \approx G_a(1 + 2Af^2).$$

Experimentally, the measurement was made by adjusting the diode to give a suitable conductance under space-charge limited conditions. This value was measured at 750 c/s and a range of radio frequencies by substitution for an I.R.C.-type MPM resistor of value 2,700 ohms.

Results for two valves are shown as a plot of $\left(\frac{G_a' - G_a}{G_a'}\right)$ 100 against frequency, giving the

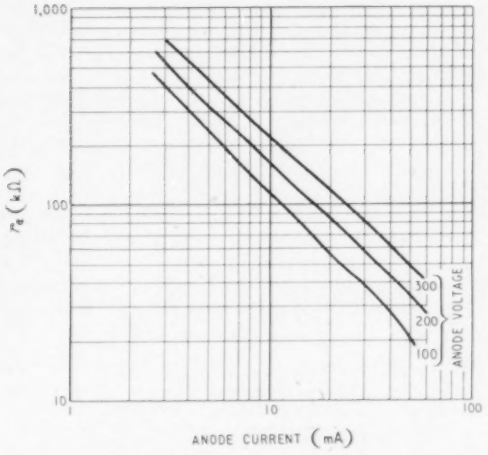


Fig. 7. *CV172 noise diode.*

percentage correction to be applied to the formula for noise output (Fig. 10). It is interesting to note that the valve giving the greater error was a special sample with electrode structure mounted lower so as to reduce the lead inductances. According to these results, the valve would be useless in the region of 400–500 Mc/s. The method of measurement is one which may be subject to error due to the conductance of the comparison resistor being frequency dependent, but more probably due to transit-time modifications of the space-charge conductance of the diode and this would have been minimized by working much closer to saturation. Any change of conductance of the comparison resistor would be expected to be in such a sense as to give an error minimizing the measured effect, but the effect of transit-time may be greater than the lead-inductance effect. It is concluded that error is no greater than on the curves. Love[6] has measured an error of 13% with a CV172 at 200 Mc/s. The

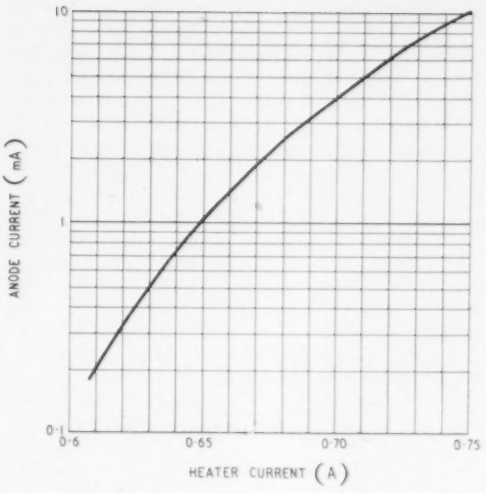


Fig. 8. *CV2171 anode-current control characteristics.*

error with the CV2171 should be much less than with the CV172.

Tests on the effects of cable mismatch showed that changes in cable length can give significant change in noise output using a value of 300 ohms for the padding resistor at the receiver input. This showed an accurate match to the cable was desirable, but to avoid errors due to changes in R_1 a half wavelength of cable is advisable if a short cable is inconvenient.

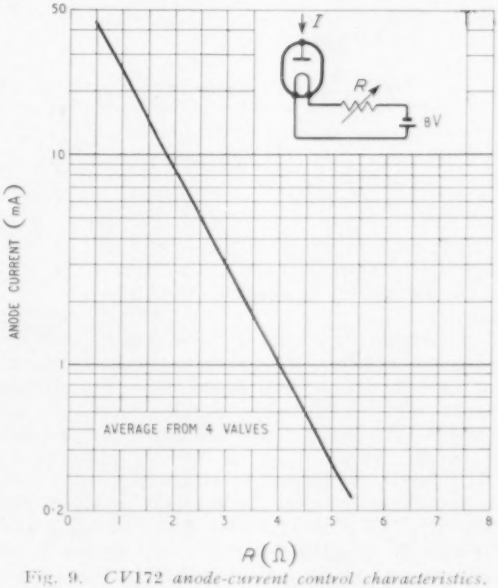


Fig. 9. *CV172 anode-current control characteristics.*

4.2. *Noise Factor*

It has been found that results over a period of years with different observers have been consistent to about $\pm$ 3% with exceptions due to traceable causes. The main sources of experimental deviations are meter errors, error in the reading of the square-law output meter (now very small), and the effects of temperature. It is advisable to check the noise-generator resistors at regular intervals.

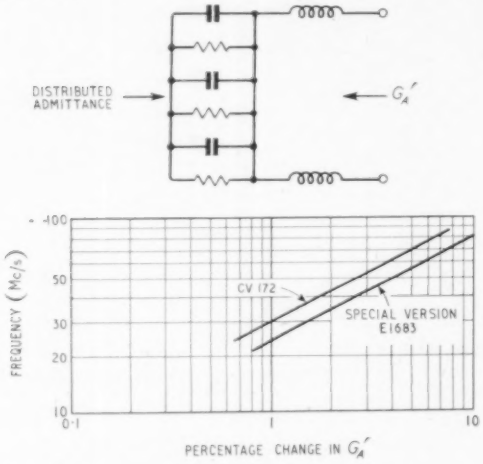


Fig. 10 (above). *CV172, measurement of transformation of diode conductance at output terminals; $G_A = 1/2700$ mhos.*

Fig. 11 (right). *Variations of mutual conductance with time (CV139).*

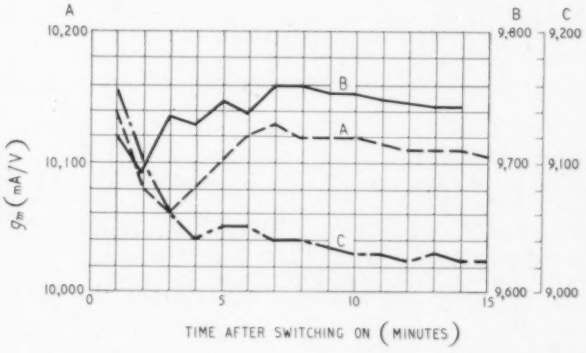
Cross-checks made in 1944 against the signal-generator method gave agreement within 2%. This is not considered reliable evidence of absolute accuracy, for the signal-generator method involves such a lot of calibration that it is not possible to avoid the suspicion of having suspended investigations of accuracy once reasonable agreement has been reached.

Cross-checks against the hot-lamp method[7] were made in 1944. Various discrepancies between the two methods were found depending upon whichever figure for lamp temperature was used, namely:—Maker's figures, calculated figures, or measured figures using an optical pyrometer. The pyrometer indicated end-cooling effects of the filament and the measurement of temperature was very difficult due to the glass background. The conclusion reached was that comparison with a noise diode enables measurement of the mean-noise temperature of the lamp filament for that particular frequency to be made.

4.3. *Shot Noise and Partition Noise*

The accuracy of these measurements has been found to be worse than expected. It is probable that real variations in valve noise are the chief cause of the experimental differences noted. In particular, when results are transformed into equivalent noise-resistance values, the value of mutual conductance is involved, and it is doubtful whether this is sufficiently stable (because of valveholder and cathode temperature effects) to justify expectation of consistency in equivalent noise resistance to better than 5%. For any careful investigation it is recommended that particular care be taken to stabilize the valve operating conditions, and to allow a period of warming up. Typical effects with a CV139 are shown in Fig. 11.

Experimentally it has been found that the deviations in measured noise resistance are as much as 5% over a period of days. Cross-checks have been made with other workers who used different methods and lower frequencies, and differences of 10 to 20% have been observed. On the whole, our results have given the lowest values, but no exhaustive investigation of the discrepancies has been made. The measurement using the substitution method gives consistent results over a short period and it is possible to check circuit errors (excluding valve lead-inductance effects) by comparing two noise diodes.



Experimental results in Tables 1, 2 and 3 show that the shot noise in modern high-slope valves is significantly greater than the theoretical value quoted by Harris.[8] This is believed to be due to the distortion of the field in the grid–cathode region with these valves, in which the grid wire is comparable with the pitch and the grid–

TABLE 1

Equivalent Noise Resistance—Triodes

CV139. $V_a = 250$ V; $I_a = 10$ mA

Valve No.	g_m (mA/V)	Equivalent Shot Noise Resistance (Ω)
A	9·60	320
B	9·33	325
C	9·21	330
D	9·55	310
E	9·75	340
F	9·70	330
G	9·55	320
H	9·75	320
Av.	9·55	325

$g_m \times R_{equ} = 3{\cdot}1$

Note:—Above samples were above average in performance (the bias required for 10 mA was about $1\frac{1}{4}$ V).

CV408. $V_a = 150$ V; $I_a = 10$ mA

Valve No.	g_m (mA/V)	Equivalent Shot Noise Resistance (Ω)
A	8·93	390
B	9·05	460
C	9·56	440
D	7·98	435
E	7·08	470
F	8·94	380
G	9·15	370
H	9·82	455
Av.	8·8	425

$g_m \times R_{equ} = 3{\cdot}7$

cathode clearance. Measurements on triodes with different operating conditions are given in Table 4. It can be seen that the approximate theoretical value[8] $2{\cdot}5/g_m$ is approached as the grid is made more positive to maintain the grid current with increased anode voltage. Similar results have been obtained for other valve types.

For measurements of partition noise, cross-checks were carried out initially by making an additional measurement with the valve triode-connected. The anode shot noise can then be calculated in three different ways, with errors due to the difficulty in reproducing operating conditions and inaccuracies in measurements of anode and screen currents. The partition noise can also be calculated for the screen and anode separately. The total spread in results for equivalent noise resistance on any one valve measured in different ways was approximately 10%.

For the CV138 and 6AK5 pentodes, it is found that partition noise is significantly lower than given by the theoretical expression

$$i^{2}_{SS} = 2e\frac{I_a I_{g2}}{I_a + I_{g2}}(1 - F^2)$$

The discrepancy is greater than would be accountable to experimental error and is probably due to the screen intercepting directly an appreciable fraction of the current from the space-charge, together with the compensating fluctuations.

5. Discussion and Conclusion

Although the noise diode has proved of enormous value in improving the reliability of noise measurements, it is still necessary to proceed with extreme care in making absolute measurements. It is the writer's experience that the more common tendency is to err on the side of optimism in measurements of noise factor for many of the less obvious errors are in that sense. For example, if slightly different answers were obtained with two different types of noise diode in the same unit, it would be the normal

TABLE 2

6AK5 *Equivalent Noise Resistance*

Pentode connected.

$V_a = 180$ V; $V_{g2} = 120$ V; $I_a = 7{\cdot}5$ mA

Valve No.	g_m (mA/V)	I_{g2} (mA)	Equivalent Shot Noise Resistance (Ω)	Equivalent Partition Noise Resistance (Ω)	
				Measured	*Theoretical*
A	4·85	3·39	1,620	930	1,480
B	4·88	2·23	750	970	1,270
C	5·26	2·48	670	780	1,180
D	4·95	2·22	820	760	1,250
E	4·98	2·42	830	815	1,260

Triode connected. $V_a = 105$ V; $I_a = 8$ mA

Valve No.	g_m (mA/V)	Equivalent Shot Noise Resistance (Ω)	$g_m \times R_{equ}$
A	5·84	840	4·9
B	6·08	570	3·45
C	6·34	560	3·55
D	6·22	570	3·55
E	6·12	590	3·60

Triode connected. $V_a = 100$ V; $I_a = 14$ mA

Valve No.	g_m (mA/V)	Equivalent Shot Noise Resistance (Ω)	$g_m \times R_{equ}$
F	7·7	470	3·6
G	8·0	410	3·3
H	7·7	505	3·9
J	7·4	470	3·5
K	8·6	400	3·4
L	7·0	450	3·1

human tendency to choose that giving the better result, yet the one giving the lower value of noise factor would probably have more error due to lead inductance. On the other hand errors in the reverse sense are more thoroughly investigated.

Certain improvements in the equipment described give reduced experimental error. In particular, recent experience with amplifiers using soldered-in valves and well-stabilized supplies has shown a very significant reduction in gain drift, thus giving much greater reproducibility in measurement.

It is considered that the methods used are superior to those described by van der Ziel,[3] chiefly because he measured resistance by a resonance method using noise. Such a method is subject to error due to the effects of induced grid noise at high frequencies and his measurement of noise temperature ratio is directly dependent on this, and involves considerable calculation after five different measurements of output reading under different input conditions.

There is need for further investigation on measurements of valve shot noise, if close agreement between different workers is to be achieved.

TABLE 3

*CV*138 *Equivalent Noise Resistance*

$V_a = V_{g2} = 250$ V; $I_a = 10$ mA

Valve No.	I_{g2} (mA)	Equivalent Shot Noise Resistance (Ω)	Equivalent Partition Noise Resistance (Ω)		g_m (mA/V)
			Measured	*Theoretical*	
A	2·63	550	420	660	7·36
B	2·69	485	430	620	7·55
C	2·65	590	380	600	7·61
D	2·69	570	410	640	7·39
E	2·65	500	410	650	7·54
F	2·70	830	480	800	6·56

$V_a = V_{g2} = 200$ V; $I_a = 8$ mA

Valve No.	I_{g2} (mA)	Equivalent Shot Noise Resistance (Ω)	Equivalent Partition Noise Resistance (Ω)		g_m (mA/V)
			Measured	*Theoretical*	
A	2·17	600	560		6·88
B	2·21	550	390		7·14
C	2·18	580	340		7·25
D	2·26	560	405		6·95
E	2·18	500	430	580	7·00
F	2·19	740	520		6·10
G	1·80	435	465		6·86
H	1·91	400	390	520	7·2
J	1·89	520	480		7·4
K	1·91	480	330	370	8·2

TABLE 4

Effect of operating conditions on relation between mutual conductance and equivalent shot noise resistance

Anode volts	Anode current (mA)	Grid volts	g_m (mA/V)	$g_m \times R_{equ}$
	*CV*139	*Sample A*		
250	10·0	Normal	10·0	3·6
100	4·2	To give	8·0	2·98
120	6·2	about	9·7	2·67
140	8·8	1-μA grid	11·0	2·68
160	11·8	current	11·7	2·69
	*CV*139	*Sample B*		3·55
113	4·0	To give	7·4	
133	5·0	about	8·2	3·51
193	11·4	1-μA grid	12·0	3·0
211	13·2	current	12·8	2·85
	6*AK*5	*Triode connected*		
105	8·0		6·3	3·55
41	3·0		4·6	3·47
61	7·0	To give	7·1	3·07
72	9·5	about	7·9	3·02
82	11·3	1-μA grid	8·3	2·96
92	14·5	current	9·1	2·95
108	19·4		9·9	2·97

The equivalent noise resistance of modern valves does not fit so closely with theory as is sometimes assumed, and the shot noise should always be checked in experiments to check noise factor theory.

Acknowledgments

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The author is grateful to his colleagues for assistance with the experiments and construction of apparatus, and particularly to Mr. A. E. Glennie, formerly of T.R.E., for his contribution to the techniques.

REFERENCES

[1] B. J. Thompson, D. O. North and W. A. Harris, "Fluctuations in Space-charge Limited Currents at Moderately High Frequencies," Part II, *R.C.A. Review*, July 1940.

[2] G. F. Valley and H. Wallmann, "Vacuum Tube Amplifiers," M.I.T. Radiation Laboratory Series, Vol. 18, Chap. 14, McGraw-Hill.

[3] A. van der Ziel, "Method of Measurement of Noise Ratios and Noise Factors," *Philips Research Reports*, October 1947, Vol. 2, No. 5.

[4] N. Houlding, "Noise Factor of Conventional V.H.F. Amplifiers," *Wireless Engineer*, November and December 1953, Vol. 30, Nos. 11 and 12, pp. 281, 299.

[5] Ref. 2. Page 714.

[6] A. W. Love, "The Behaviour of a Diode Noise Generator at V.H.F." *J. Instn Engrs, Aust.*, April/May 1948, Vol. 20, No. 4-5.

Note:—In this paper the effect of anode voltage on noise output is also investigated. The results are not explained but appear to be due to the finite slope resistance shunting the 1000-ohm load resistance used.

[7] E. H. Ullrich and D. C. Rogers, "An Absolute Method of Measurement of Receiver Noise Factor," *J.Instn elect. Engrs*, 1946, Vol. 93, Part IIIA.

[8] B. J. Thompson, D. O. North and W. A. Harris, Ref. 1, Part V, *R.C.A. Review*, April 1941, p. 514.

CORRESPONDENCE

Letters to the Editor on technical subjects are always welcome. In publishing such communications the Editors do not necessarily endorse any technical or general statements which they may contain.

Microwave Wide-Angle Scanner

SIR,—With reference to the paper of this title by J. Brown which appeared in the October issue of *Wireless Engineer*, it is a coincidence that I submitted an article on the same subject to the *Journal of Applied Physics* in August 1953. This article will be published in their March or April 1954 issue under the title of "Modified Luneberg Lens."

The following comments on the subject may be of interest to your readers:—

I have used the corpuscular theory of light (Hamiltonian Optics) and derived equation (24) of Brown's article and also was able to arrive at an expression for the index of refraction which is:

$$n = \frac{1}{a}\sqrt{a^2 + b^2 - r^2}$$

This is a simple algebraic solution of the lens treated by Brown. (See Fig. 1 of page 250 of your October issue.) This solution utilizes the full aperture $2a$ of the lens; n is continuous and there are no discontinuities in dn/dr or higher derivatives.

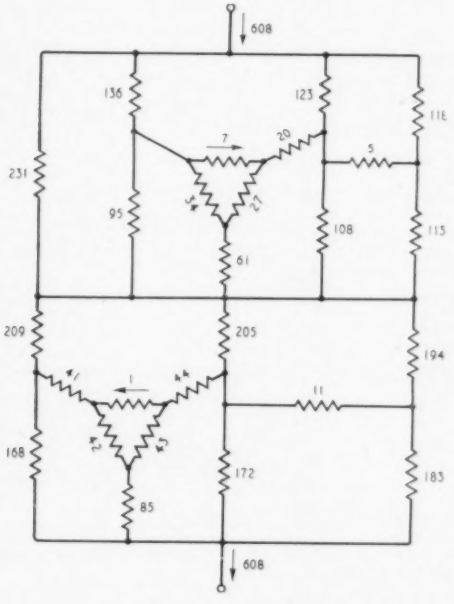
ALFRED S. GUTMAN.

A.F. Cambridge Research Center,
Brookline, Mass., U.S.A.
24th November 1953.

A Circuit Problem

SIR,—May I propose an electric circuit problem of an unusual kind, and which is, I believe, of exceptional difficulty.

It is required to connect the smallest possible number of 1-ohm resistors into a two-terminal network in such a way that no two resistors carry equal currents and so that the resistance of the combination is 1 ohm. The trivial case of one single resistor is excluded.



In the accompanying figure, which presents a first attempt on the problem, twenty-six 1-ohm resistors are shown connected together into a network whose resistance is 1 ohm. The relative currents are marked alongside the resistors and it will be seen that no two resistors carry the same current.

Before this circuit can be accepted as the solution of the problem it must be proved that no solution exists using fewer elements. At least one further network must exist which satisfies the conditions equally well; viz; the dual of the first, but it does not use fewer elements. Failing the desired proof, it would nevertheless be an advance if someone could produce a circuit using fewer elements or show that one must exist.

This curious problem is related to, but more general than, the problem of dividing a square into smaller squares.[1]

R. N. BRACEWELL.

Commonwealth Scientific and
Industrial Research Organisation,
Chippendale, N.S.W., Australia.
19th November 1953.

REFERENCE

[1] R. L. Brooks, C. A. B. Smith, A. H. Stone and W. T. Tutte, "The Dissection of Rectangles into Squares", *Duke Math. J.*, 1940, Vol. 7, p. 312.

NEW BOOKS

Flywheel Synchronization of Saw-Tooth Generators

By P. A. NEETESON. Television Receiver Design, Monograph 2. Philips Technical Library. Electronic Valves, Vol. VIIIB. Pp. 156 + vi. Cleaver-Hume Press, Ltd., 42a South Audley Street, London, W.1. Price 21s.

Flywheel sync is now used in virtually all American television receivers and in many, if not most, British fringe-area models. In spite of this, the literature on it is confined to a few serious papers and rather elementary descriptive articles. The appearance of a book on the subject is, therefore, a matter of importance to television receiver designers.

The first 41 of the 156 pages comprise, in five out of the six chapters of the book, what is doubtless intended as a refresher course on time-bases. Most readers will be familiar with the material in them.

Chapter 6 has 101 pages and is entitled Flywheel Synchronization. It is a curious and not very happy arrangement to devote five chapters to subsidiary matters and to make the heart of the book one chapter which is so long that it has to be broken up into sections, sub-sections and sub-sub-sections.

This chapter has four main sections numbered 6.1 to 6.4, of which the first two are an introduction and an explanation of the principle of flywheel synchronization, all in three pages. Section 6.3 is headed "Flywheel Action of Resonant Circuits" and 6.4 is "Automatic Phase Control". Each of them is further divided and, surely, absurdity is reached when 6.4.1 is divided into 6.4.1.1, although there is no 6.4.1.2, and 6.4.1.1 is divided into

6.4.1.1.1 to 6.4.1.1.9. Numbers like these are just a nuisance to the reader and are too difficult to remember to be of any use for reference purposes; anyway, what are page numbers for?

When the reader has penetrated all these barriers, he will find that Section 6.3 on the flywheel action of resonant circuits is a very important one which contains a great deal of useful information. It is mainly an analysis of a resonant circuit excited by regularly-recurring very-narrow pulses. The conditions which exist when the resonant circuit is the same as, is slightly different from, and is a multiple of an integer-plus-one-half of, the recurrence frequency are very fully investigated mathematically. The explanations are too brief, however, and the reader has to work harder than he should to make use of the results. The investigation leads to a short description of an unusual form of flywheel circuit. It is one in which the sync pulses energize a tuned circuit to produce a near sine-wave output which is then squared and differentiated to develop new pulses which are used to lock the line time-base. No feedback is involved.

Section 6.4 covers the more usual kind of flywheel circuit in which the sync pulses are compared in a phase discriminator with pulses derived from the time-base and an error signal is produced to control the time-base. The form of phase discriminator is not the usual one, being a multi-electrode valve which passes a current depending on the overlap of pulses applied to two different grids. It is, perhaps, a convenient one for explanatory purposes, but the common two-diode circuit should have been fully treated instead of receiving only one page of very elementary description.

A very great deal of the space in this section is devoted to an analysis of the multivibrator, both as a square-wave and as a narrow-pulse generator. One's first reaction is that much of this is irrelevant but it is a wrong reaction, for the end of it all is to show how the frequency of a multivibrator is dependent on a grid-control voltage, a matter which is at the root of automatic phase control.

The system is basically of the negative-feedback type and both static and transient conditions are investigated in considerable detail. It is shown that a disturbance, such as that occuring during the frame pulses, can produce an oscillatory response which manifests itself as a distorted edge to the picture. There is no adequate discussion of the requirements for avoiding such effects. One is left with rather a gloomy impression of the performance of such circuits and, if one did not know that they were so widely used and performed so well in practice, one would feel that they were quite unsuitable for television!

The book has many faults in the presentation of the material and in a lack of clarity of exposition. Too often the author explains in unnecessary detail the things everyone knows and treats quite briefly, almost in passing, things that few readers will be familiar with. In spite of this, however, it is redeemed by the importance of the subject. Much of the material has either not been previously published or is not in any readily accessible media. The book will give the reader a great deal of information about flywheel sync circuits and is a necessary book for the designer. He will, however, have to work hard to extract the information he wants from it.

W. T. C.

F.B.I. Register of British Manufacturers 1954

26th Edition. Pp. 952. Published for the Federation of British Industries by Kelly's Directories Ltd., and Iliffe & Sons, Ltd., Dorset House, Stamford Street, London, S.E.1. Price 42s. (including postage).

Cours sur les Ondes Ultra-Courtes

By Y. Place. Pp. 185. Editions Eyrolles, 61 Boulevard Saint-Germain, Paris Ve, France. Price 1,300 fr.

I.E.E. MEETINGS

13th January. "A Single-Sideband Controlled-Carrier System for Aircraft Communication", by G. W. Barnes, B.Sc.

18th January. "The Role of the Consulting Engineer." Discussion to be opened by T. G. N. Haldane, M.A.

19th January. "High-Sensitivity Wattmeters." Discussion to be opened by A. H. M. Arnold, Ph.D., D.Eng.

20th January. "The Teaching of Magnetic Materials." Discussion to be opened by Professor F. Brailsford, Ph.D., B.Sc.(Eng.) at 6 p.m.

25th January. "Should Sound Broadcasting of the Future be entirely in the V.H.F. Band?" Discussion to be opened by A. J. Biggs, Ph.D., B.Sc.

These meetings will be held at the Institution of Electrical Engineers, Savoy Place, Victoria Embankment London, W.C.2, and will commence at 5.30 except where otherwise stated.

STANDARD-FREQUENCY TRANSMISSIONS

(Communication from the National Physical Laboratory)

Values for November 1953

Date 1953 Nov.	Frequency deviation from nominal: parts in 10^8		Lead of MSF impulses on GBR 1000 G.M.T. time signal in milliseconds
	MSF 60 kc/s 1429-1530 G.M.T.	Droitwich 200 kc/s 1030 G.M.T.	
1	N.M.	0	N.M.
2	− 0·8	+ 2	− 43·4
3*	—	+ 1	—
4	− 0·8	+ 1	− 45·4
5	− 0·7	0	− 47·2
6	− 0·7	+ 1	− 47·6
7	− 0·6	0	N.M.
8	− 0·6	+ 1	N.M.
9	− 0·6	+ 1	− 50·7
10	− 0·5	+ 2	− 52·0
11	− 0·5	+ 1	− 52·8
12	− 0·5	+ 1	− 53·2
13	− 0·5	+ 1	− 53·7
14	− 0·5	+ 2	N.M.
15	− 0·4	+ 2	N.M.
16	− 0·4	+ 2	− 54·6
17	− 0·4	+ 1	− 56·1
18	− 0·5	+ 2	− 56·6
19	− 0·5	+ 3	− 57·2
20	− 0·4	+ 2	− 57·3
21	− 0·5	+ 3	N.M.
22	− 0·5	+ 1	N.M.
23*	—	+ 1	—
24*	—	+ 2	—
25*	—	+ 2	—
26	− 0·3	+ 2	− 60·1
27	N.M.	+ 3	N.M.
28	− 0·3	+ 2	N.M.
29	N.M.	+ 3	N.M.
30	− 0·2	+ 2	− 64·3

The values are based on astronomical data available on 1st December 1953.

The transmitter employed for the 60-kc/s signal is sometimes required for another service.

N.M. = Not Measured.

* = No Transmission.

POLYTAGS... lead-through and stand-off insulators

Polytetrafluoroethylene (P.T.F.E.) is an outstanding insulator. It is tough, durable and will not crack or arc. Its dielectric properties are substantially constant over a frequency range of 60 c.p.s. to at least 300 Mc.p.s. and are unaffected by temperature changes between minus 100°C. and plus 288°C. It has zero moisture absorption and is water repellent. It is, therefore, a most suitable material for stand-off and feed-through insulator terminals and has been chosen by Ediswan for this purpose. Ediswan Polytags are available in five types as illustrated below.

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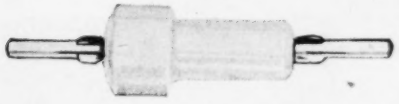
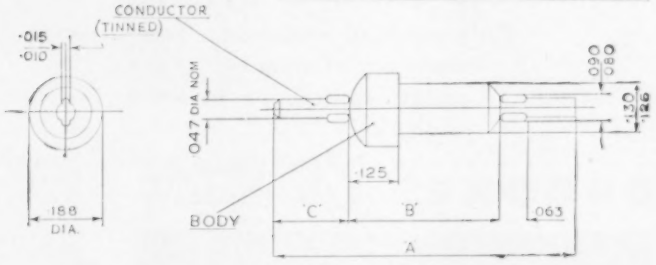
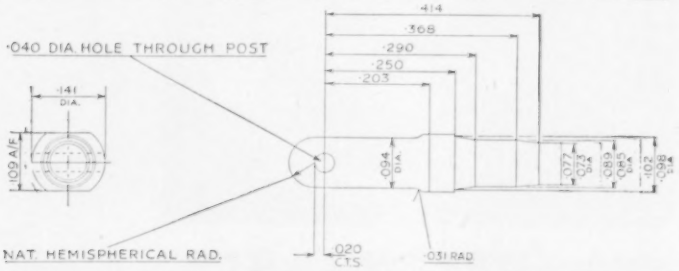


PT 3 & 4. Stand-off

	A	B	C
PT 1	.750	.375	.188
PT 2	.875	.500	.188
PT 3	.563	.375	—
PT 4	.688	.500	—



PT 5. Component mounting



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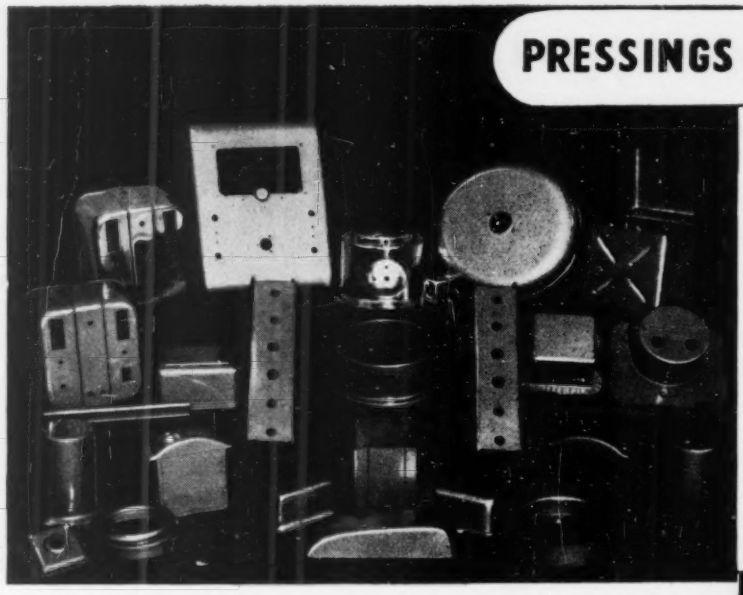

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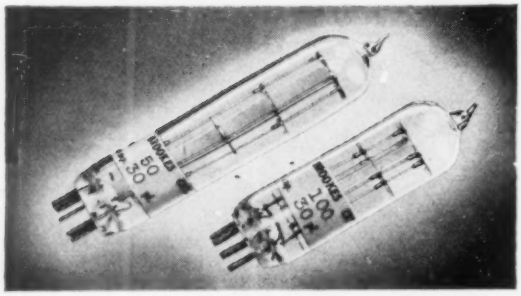

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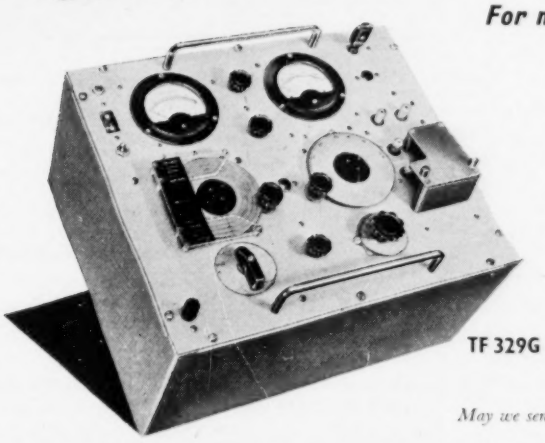

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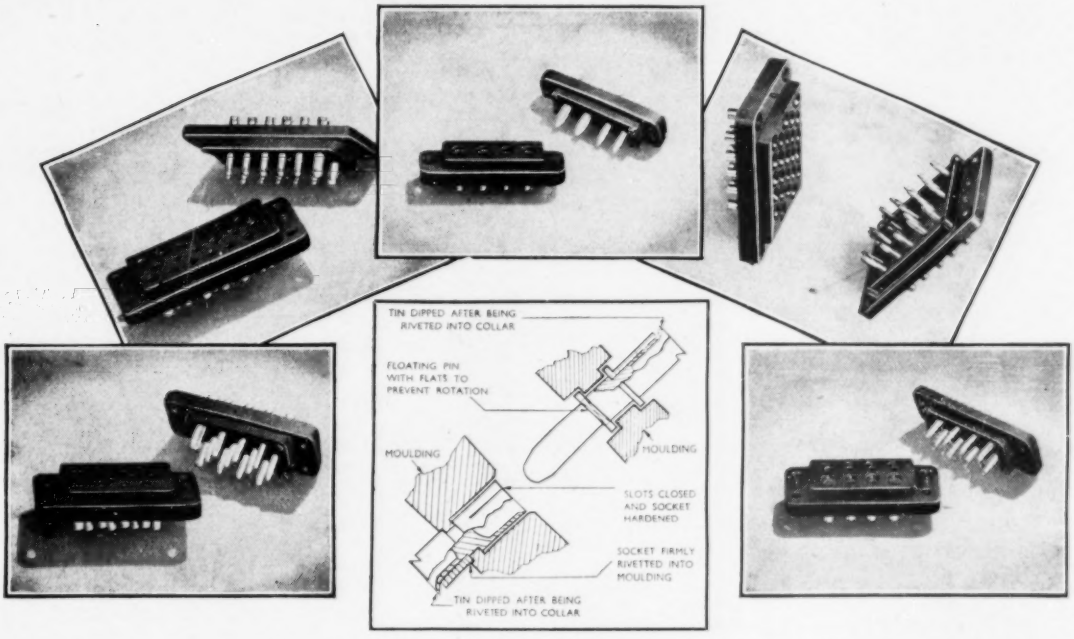
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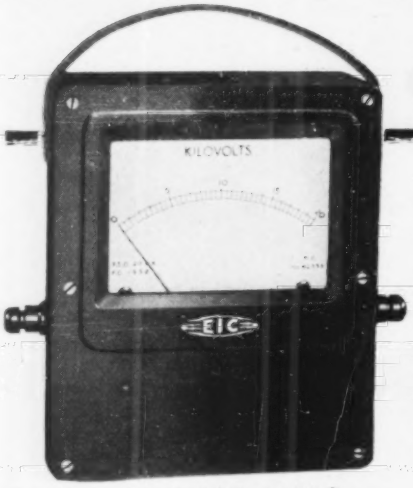

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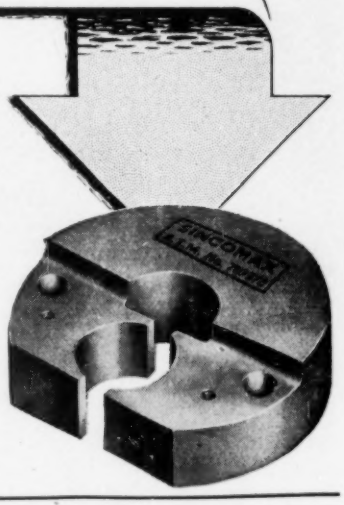

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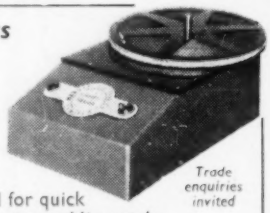

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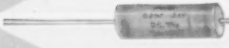

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